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RF power amplifier linearisation using predistortion techniques.

RF Power Amplifier Linearisation Using Predistortion Techniques

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A thesis submitted to the University of Bristol in accordance with the requirements for the degree of Doctor of Philosophy in the Faculty of Engineering, Department of Electrical and Electronic Engineering.

Abstract

Linear amplifiers are becoming increasingly important as the drive for more spectrally efficient systems continues. As the number of people wishing to use mobile communication systems increases then broadband multi-user systems will become necessary for future communications needs. This thesis presents a new broadband polynomial predistortion system that is capable of linearising an amplifier over a decade of frequency. Various techniques for linearising amplifiers are available and the most successful narrowband and broadband techniques are explained. Particular emphasis has been placed on the development of a system for broadband applications, which permits access by a large number of users. The feedforward technique provides an excellent improvement in linearity but results in poorer efficiency when compared with the original amplifier. Predistortion has a much lower efficiency penalty than feedforward but will offer only modest improvements in linearity. If a predistorter could be developed which would give a modest improvement in linearity for a minimal reduction in efficiency then the efficiency of the overall system would be increased. Another advantage of this approach is the potential reduction in the number of loops within the feedforward amplifier that will reduce complexity and cost while retaining good linearity performance for a hybrid amplifier system. A mathematical analysis of a polynomial predistorter is presented which shows the potential performance benefits of predistortion and the required gain and phase matching. A practical polynomial predistortion system has been developed and investigated. The system contains a new type of polynomial predistorter with significantly better performance than systems previously published. It has been shown that for good broadband performance the predistorter needs to be carefully designed taking account of the electrical delay through the system. The system has been developed to operate over a wide frequency range. With careful design the system has been developed for narrowband, mediumband and broadband applications. The polynomial predistorter has also been developed to provide improvements in the linearity of a class C amplifier. This thesis also shows that this technique can improve the efficiency of a feedforward amplifier by 11%, which results in a hybrid amplifier with an efficiency of 38%. This thesis clearly shows that this technique provides a potentially very high performance polynomial predistorter.

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AUTHOR'S DECLARATION

Unless otherwise acknowledged, the content of this thesis is the original and sole work of the author. No portion of the work in this thesis has been submitted by the author in support of an application for any other degree or qualification, at this or any other university or institute of learning. The views expressed in this thesis are those of the author, and not necessarily those of the University of Bristol.

Kevin Andrew Morris

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ABBREVIATIONS

ABP	Adaptive Baseband Predistortion
bps	Bits per second
BJT	Bipolar Junction Transistor
BPSK	Binary Phase-Shift Keying
BS	Base Station
BTS	Base Transmission Site
CALLUM	Combined Analogue Locked-Loop Universal Modulator
CDMA	Code Division Multiple Access
dB	Decibel
dBc	Decibels with respect to carrier
dBm	Decibel relative to one milliwatt
DSP	Digital Signal Processing
EE&R	Envelope Elimination and Restoration
GaAs	Gallium Arsenide
GMSK	Gaussian Minimum-Shift Keying
GRAN	Graduated Radio Access Networks
GSM	Global System for Mobile Communications
GSM EDGE	GSM Enhanced data rates using optimised Modulation
HF	High Frequency
IF	Intermediate Frequency
IMD	Intermodulation Distortion
IMP	Intermodulation Product
IP	Intercept Point
I&Q	In Phase and Quadrature
kbps	Kilo bits per second
LINC	Linear Amplification using Nonlinear Components
LIST	Linear Amplification using Sampling Techniques

LO	Local Oscillator
LUT	Look-Up Table
Mbps	Million bits per second
MCPA	Multi-Carrier Power Amplifiers
MS	Mobile Station
8- OPSK	8 Offset Phase Shift Keying
QPSK	Quadrature Phase Shift Keying
RF	Radio Frequency
SSB	Single Side Band
SMS	Short Message Service
TETRA	Trans-European Trunked Radio
TWT	Travelling-Wave Tube
UHF	Ultra High Frequency
UMTS	Universal Mobile Telecommunications System
UTRA	Universal Mobile Telecommunications System Terrestrial Radio Access
VHF	Very High Frequency

PUBLICATIONS

Kevin Morris and Peter Kenington, Power Amplifier Linearisation Using Predistortion Techniques. In IEE Colloquium on RF and Microwave Components for Communications Systems, pp. 6/1 – 6/6, April 1997.

Kevin Morris and Peter Kenington, A Broadband Linear Power Amplifier for Software Radio Applications. In Proceedings of the 48th IEEE Vehicular Technology Conference, pp. 2150 – 2154, Ottawa May 1998.

Chapter 1
Introduction

This chapter introduces trends within modern communications systems and the requirements for amplification techniques beyond the current state of the art.

1 Introduction

1.1 The Evolution of Communications Systems

Communications systems are moving into a new era with GSM now firmly established as the current standard. Operators and equipment manufacturers are looking to extend the capabilities of the systems they offer to maintain market momentum. It is now firmly established that global roaming and increased data capacity are aims that will need to be met. The current ETSI debate surrounding so-called 3rd generation systems such as UMTS [1,2,3], GSM+, GSM EDGE [4] and CDMA2000 is gathering pace with the big players vying for position. The main thrust of all these new techniques is the desire to increase data rates to levels that are significantly above the rates that are currently achievable by GSM. The aim of this drive is to allow people to access a broad range of data sources and in the medium term to deliver World Wide Web access to hand held mobile terminals. This will inevitably place demands on the system hardware employed.

1.2 The Software Radio Concept

These standards operate with various air interfaces at a variety of frequencies and over a range of bandwidths, creating differing amplifier performance requirements. So system operators now require hardware that is capable of supporting various standards with minimal modification. Current research is aimed at the evolution of the software radio concept [5,6] which would contain hardware capable of supporting multiple standards with a reconfigurable software base. The concept of a software radio with global roaming capability is shown in figure 1.1.

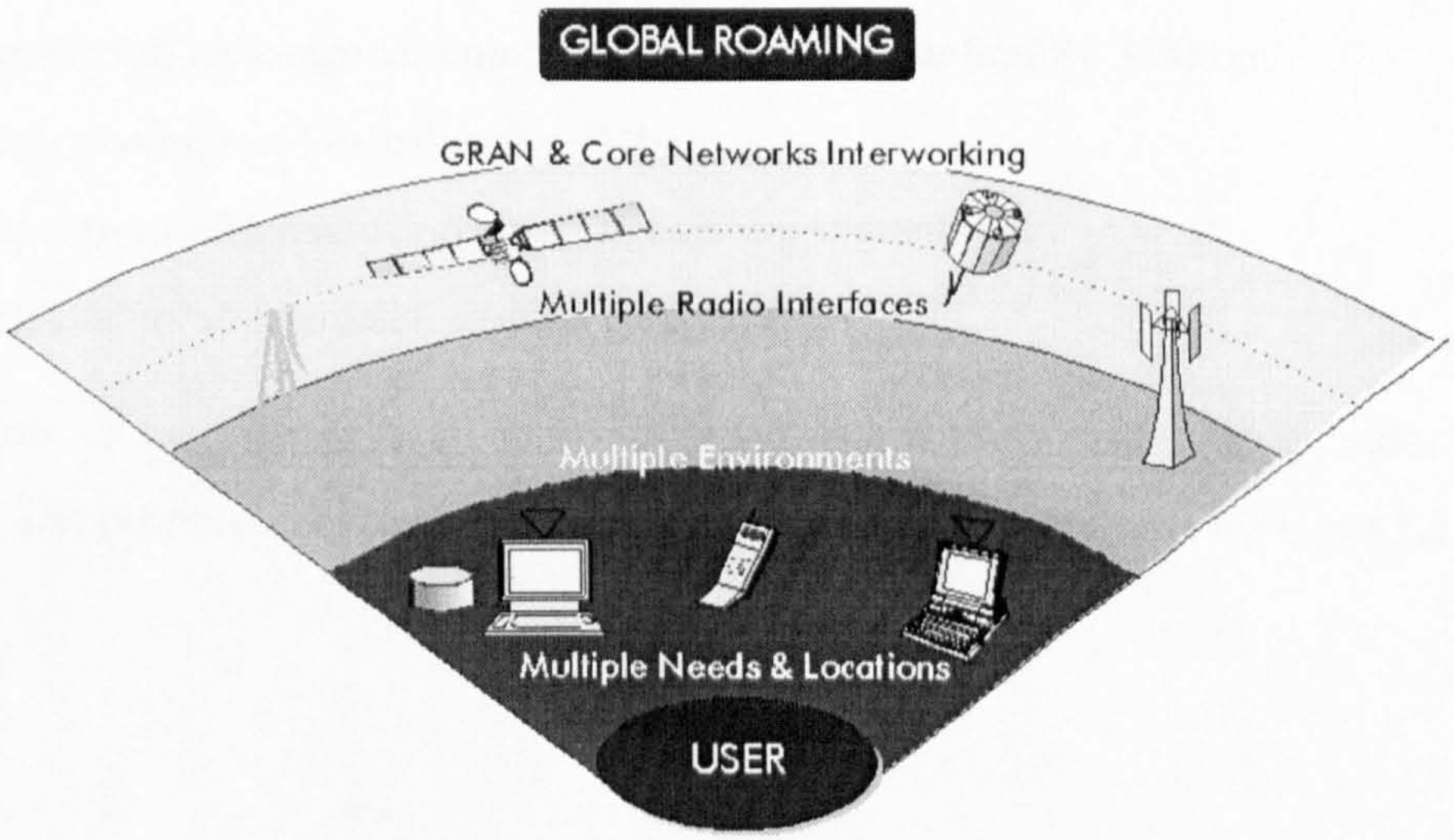

Figure 1.1 The Global Roaming Software Radio

Current mobile system hardware has partially achieved this aim by using multiple parallel radio chains (dual-band and tri-band systems) driven by a single baseband software system which allows a level of global roaming in the case of GSM. The use of multiple R.F. chains has an inevitable impact on the efficiency of the radio system however. For example a single band mobile has a standby time of 7 days and a use time of 4 hours, whereas a dual band mobile has a standby time of 5 days and a use time of 3 hours[1]. With the customer expectation of these long standby and use times the efficiency and power consumption of the radio hardware is very much to the fore of discussion and research direction in terms of amplifier design.

1.2.1 System Bandwidth Requirements

Second Generation systems operate with a channel spacing of 200kHz and a total occupied bandwidth of 25MHz either side of the frequency allocation for the uplink and the downlink. GSM also operates using a constant envelope modulation scheme known as GMSK. This relaxes the amplifier linearity requirements significantly. Current second-generation systems such as GSM are limited in terms of the achievable data rate that maybe transmitted, although the next evolution of GSM known as GSM Phase 2+ (GSM EDGE etc.) is addressing these issues. Current GSM systems support a data rate of 9.6kbps. This allows the system to support voice and limited data services such as SMS and e-mail. The next phase of GSM is capable of supporting data rates of up to 182.4kbps. This upper data rate is likely to be the maximum rate that will be supported by the GSM system.

The development of UMTS has been fuelled by the need to provide higher data rates than are currently possible and with the desire to create a global communications standard. The aims of UMTS are:

- To support full coverage and mobility for 144kbps, preferably 384kbps
- Limited coverage and mobility for 2Mbps
- High spectrum efficiency compared to existing systems
- High flexibility to introduce new services.

UMTS aims to provide a broad range of services currently unavailable under the GSM standard. The proposed range of services offered by UMTS are shown in figure 1.2.

[1] The figures are based on single band and dual band Nokia mobile phones. Data obtained September 99 from http://www.uk.orange.net

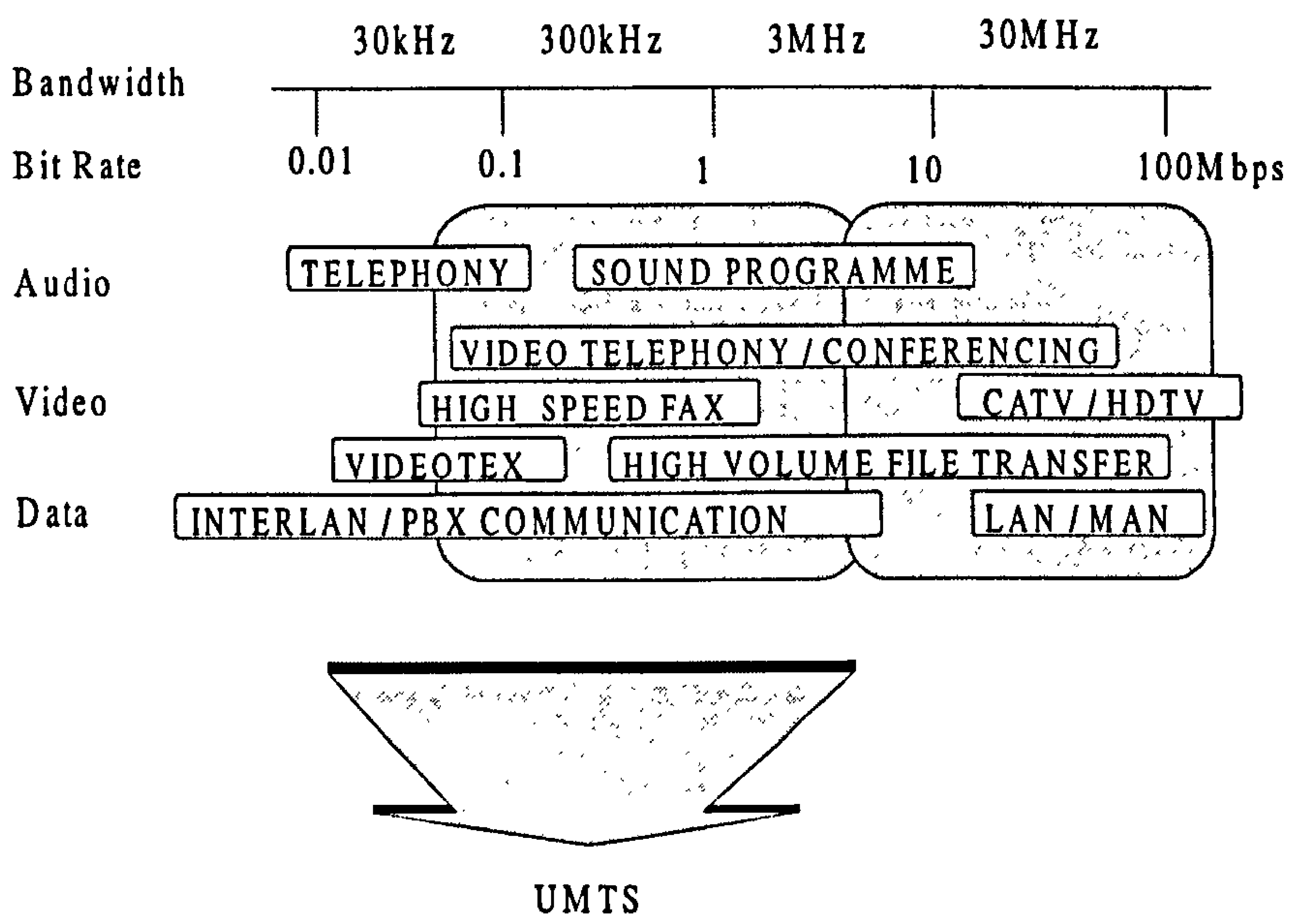

Figure 1.2 Bandwidth Requirements of UMTS

In order for UMTS to achieve the required improvements in data rate over GSM the UMTS system has been designed to operate over a channel bandwidth of 5MHz using QPSK and BPSK modulation schemes, this has implications on the required power amplifier performance.

A table of the various amplifier requirements for GSM, GSM EDGE [4] and UMTS [7,8,9,10] are shown in Table 1.1.

Standard	Modulation Scheme	Channel Bandwidth kHz	BTS/BS[2] 1st Adjacent Channel Power	BTS/BS[2] 2nd Adjacent Channel Power	MS 1st Adjacent Channel Power	MS 2nd Adjacent Channel Power
GSM	GMSK	200	-30dBc	-60dBc	-30dBc	-60dBc
GSM-EDGE	8 OPSK	200	-30dBc	-60dBc	-30dBc	-60dBc
UTRA-FDD	QPSK	5000	-45dBc	-50dBc	-33dBc	-43dBc
UTRA-TDD	QPSK	5000	-45dBc	-55dBc	-33dBc	-43dBc

Table 1.1 Amplifier Linearity Requirements of Modern Communication Systems

[2] BTS used to denote GSM/GSM EDGE basestation. BS used to denote UTRA basestation.

1.3 Basestation R.F. Power Amplification

With a mobile network there are two basic methods of providing the necessary amplification of the user channels. The channels maybe individually power amplified and then power combined. Alternatively the channels are up-converted, power combined and then amplified. These two approaches are shown in figures 1.3 and 1.4 respectively.

The approach adopted in figure 1.3 has the advantage of using narrowband power amplifiers which are easier and cheaper to produce, however this approach requires high power combining which results in a significant loss in the total available output power of the amplifiers. There is also the associated additional heating effect from this loss in power that cannot be transmitted over the air.

The broadband amplifier illustrated in figure 1.4 overcomes the losses associated with high power combining by modulating and upconverting the baseband signal at low power. This signal is then combined before amplification. This low power combining results in minimal loss of power and minimal associated heat losses. The signal is then broadband power amplified, typically over several 10's of megahertz at a carrier frequency anywhere between 900-2200MHz depending on the particular requirements of the operator and the relevant countries regulating authority.

So there is a need for amplification systems which are capable of operating over significant bandwidths at high levels of efficiency. These amplifiers are also required to support multiple standards and operating frequencies. Therefore there is a need for broadband high efficiency, multi-standard R.F. power amplifiers within modern communications systems. This thesis is intended to investigate amplifier linearisation techniques that are suitable for use in 2nd and 3rd generation reconfigurable basestations.

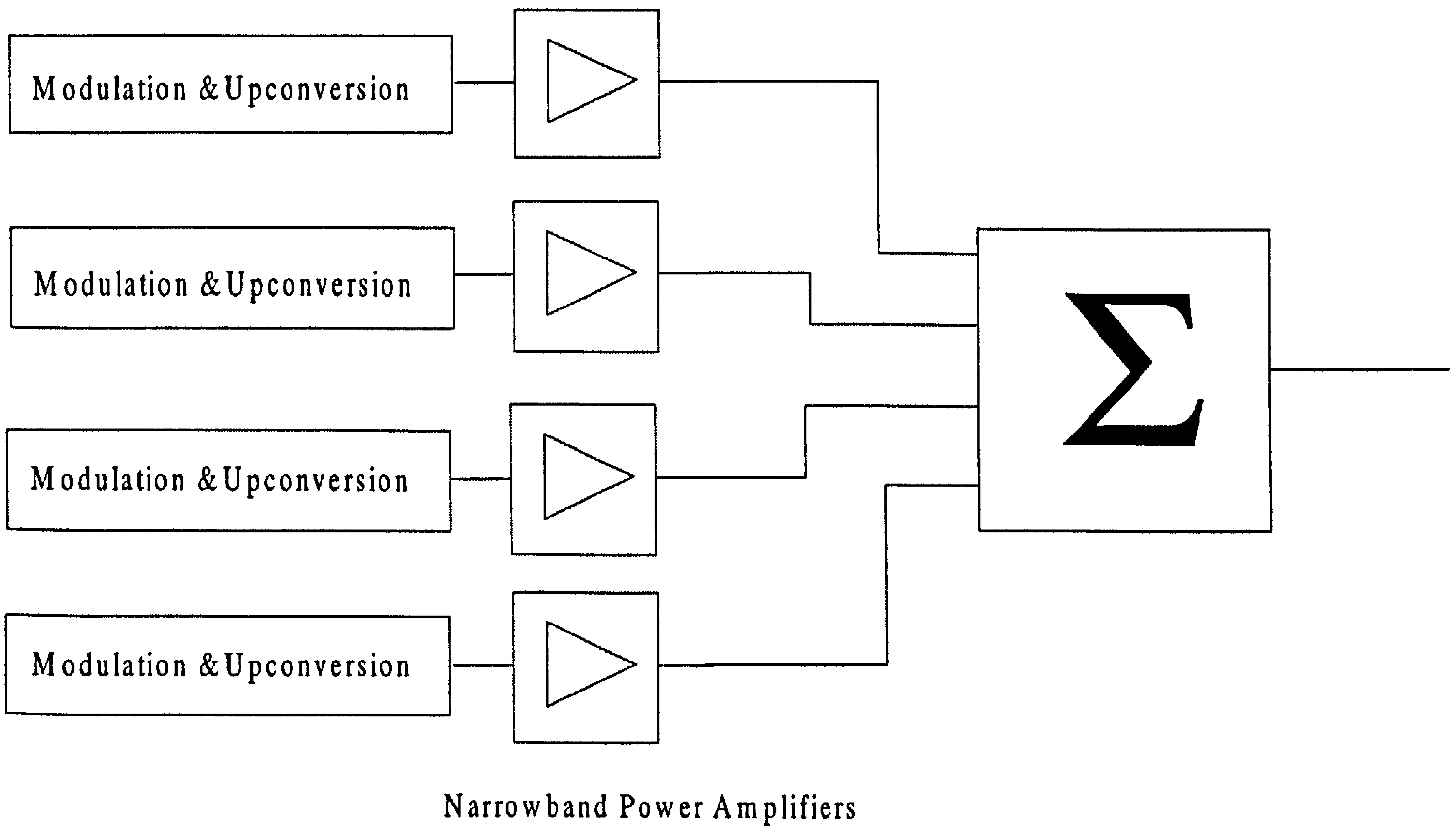

Figure 1.3 Narrowband Basestation Amplifier

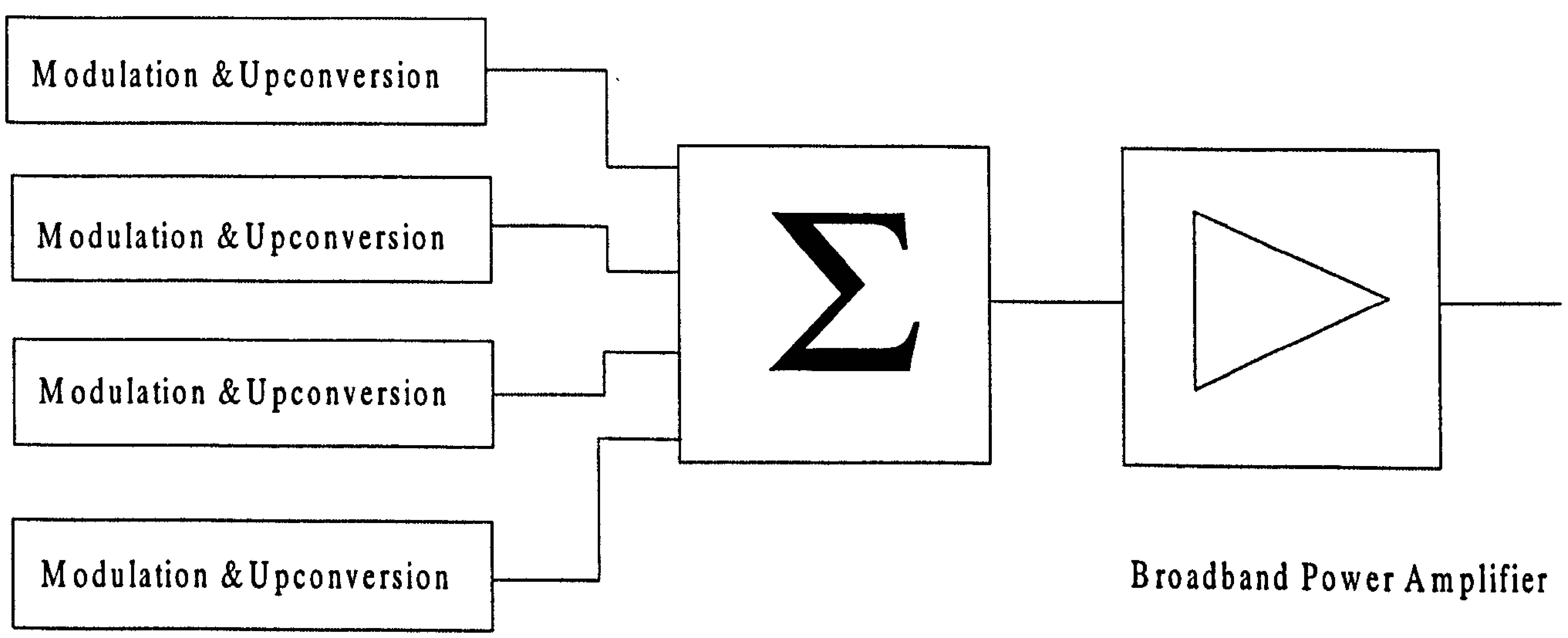

Figure 1.4 Broadband Basestation Amplifier

1.4 Structure of Thesis

Chapter 2 of this thesis covers the background to this piece of work, the areas covered include the way amplifiers are characterised, the conventional classes of amplifier and the methods which are used, or have been proposed for the linearisation of power amplifiers.

When considering the method used to linearise an amplifier it has been stated that efficiency is of prime importance when basestation and mobile power amplifiers are considered. For this reason the *R.F. predistortion technique* has been chosen for further investigation. Predistortion has been proposed previously as a linearisation technique but the majority of these systems use baseband predistortion techniques. R. F. predistortion has been chosen for its potential application within broadband systems. One form of R.F. predistortion is polynomial predistortion. Chapter 3 covers the mathematical analysis of a 3rd order polynomial predistortion system that is suitable for the broadband linearisation of a power amplifier.

Chapter 4 introduces a new form of broadband polynomial predistorter for use within a basestation or mobile amplifier system. The predistortion system is examined in detail in order to characterise its performance in terms of the possible achievable bandwidth, the achievable improvements in amplifier performance. The system's tolerance to gain and phase errors is investigated and comparisons are made with the theoretical performance obtained form the analysis within chapter 3. The effect changes in input have on system performance and the system's transfer characteristics are also investigated.

Chapter 5 introduces the possibility of utilising a polynomial predistortion system to linearise highly non-linear power amplifiers. A practical realisation of a polynomial predistortion system is presented. Investigations are presented which show the linearity that is achievable and the systems tolerance to errors.

Chapter 6 provides a summary of the thesis and from this summary some conclusions are drawn about the work carried out within the preceding chapters. This chapter also contains suggestions for future work within the area of transmitter linearisation.

References

[1] J. Rapeli, UMTS: Targets, system concept and standardisation in a global framework, IEEE Personal Communications Magazine, pp. 20 – 28, February 1995.

[2] ETSI SMG2, Evaluation Report for ETSI UMTS Terrestrial Radio Access (UTRA) ITU-R RTT Candidate, September 1998.

[3] 3rd Generation Partnership Project (3GPP); Technical Specification Group (TSG) RAN Working Group 4 (WG4); RF Parameters in Support of Radio Resource Management, TS 25.103 V2.0.0, October 1999.

[4] European Telecommunications Standards Institute, GSM/GSM EDGE Digital Cellular Telecommunications System (Phase 2+), Radio Transmission and Reception (GSM 05.05 version 7.1.1 Release 1998), ETSI EN 300 910 V7.1.1 (1999-12)

[5] J. Mitola, The software radio architecture, IEEE Communications Magazine, pp. 26 – 38, May 1995.

[6] J.M. Pereira, Beyond Software Radio, towards Re-configurability across the whole System and across Networks, VTC 1999 Fall, 50th Vehicular Technology Conference, pp. 2815 – 2818, September 1999.

[7] 3rd Generation Partnership Project (3GPP); Technical Specification Group (TSG) RAN WG4; UTRA (BS) FDD; Radio transmission and Reception, TS 25.104 V3.0.0, October 1999.

[8] 3rd Generation Partnership Project (3GPP); Technical Specification Group (TSG) RAN WG4; UE FDD; Radio transmission and Reception (FDD), TS 25.101 V3.0.0, October 1999.

[9] 3rd Generation Partnership Project (3GPP); Technical Specification Group (TSG) RAN WG4; UTRA (BS) TDD; Radio transmission and Reception, TS 25.105 V3.0.0, October 1999.

[10] 3rd Generation Partnership Project (3GPP); Technical Specification Group (TSG) RAN WG4; UTRA (UE) TDD; Radio transmission and Reception, TS 25.102 V3.0.0, October 1999.

Chapter 2
Amplifier Characteristics and Linearisation Techniques

This chapter discusses how an amplifier would ideally behave and how amplifiers behave in reality. The chapter explains how amplifiers are characterised and the various tests of linearity that are employed to measure amplifier performance practically. The chapter also carries out a review of current and future methods of amplifier linearisation which have been considered in the literature.

2 Amplifier Characteristics and Linearisation Techniques

2.1 Introduction

The issue of linear efficient broadband amplification is one of the essential requirements of any modern radio communication system. This chapter will cover the issues associated with R.F. power amplifiers. In an ideal world an amplifier would provide distortion free, 100% efficient, multi-frequency, broadband amplification using a single amplification module. Currently this situation has not been realised and so techniques have been developed to describe how an amplifier behaves in practice. Amplifier characterisation is essential if the effect that the amplifier has on the signal is to be quantified. This chapter describes how amplifiers are currently characterised and then discusses the operation of the classic amplifier classes and finally carries out a review of amplifier linearisation methods.

2.2 Amplifier Characteristics

2.2.1 Amplifier Distortion

An ideal amplifier would produce an output signal $y(t)$, which was a delayed and magnified replica of the input signal $x(t)$. Assuming a frequency flat group delay this can be expressed mathematically as

$$y(t) = Kx(t - t_0) \tag{2.1}$$

which implies that the amplifier transfer function $H(j\omega)$ must be

$$H(j\omega) = Ke^{-j\omega t_0} \tag{2.2}$$

This means that any deviation from the constant amplitude K and the constant negative phase shift $-j\omega t_0$ will result in amplitude and phase distortion as a function of frequency [1]. Besides this type of distortion, other deviations can occur if the amplifier possesses non-linear elements. In this case it is not possible to describe the amplifier by a single transfer function. Instead the output is often expressed as a non-linear function of the input, that is, $y(t) = T[x(t)]$ this is shown in figure 2.1. Signal distortion resulting from this type of deviation is often called "non-linear distortion".

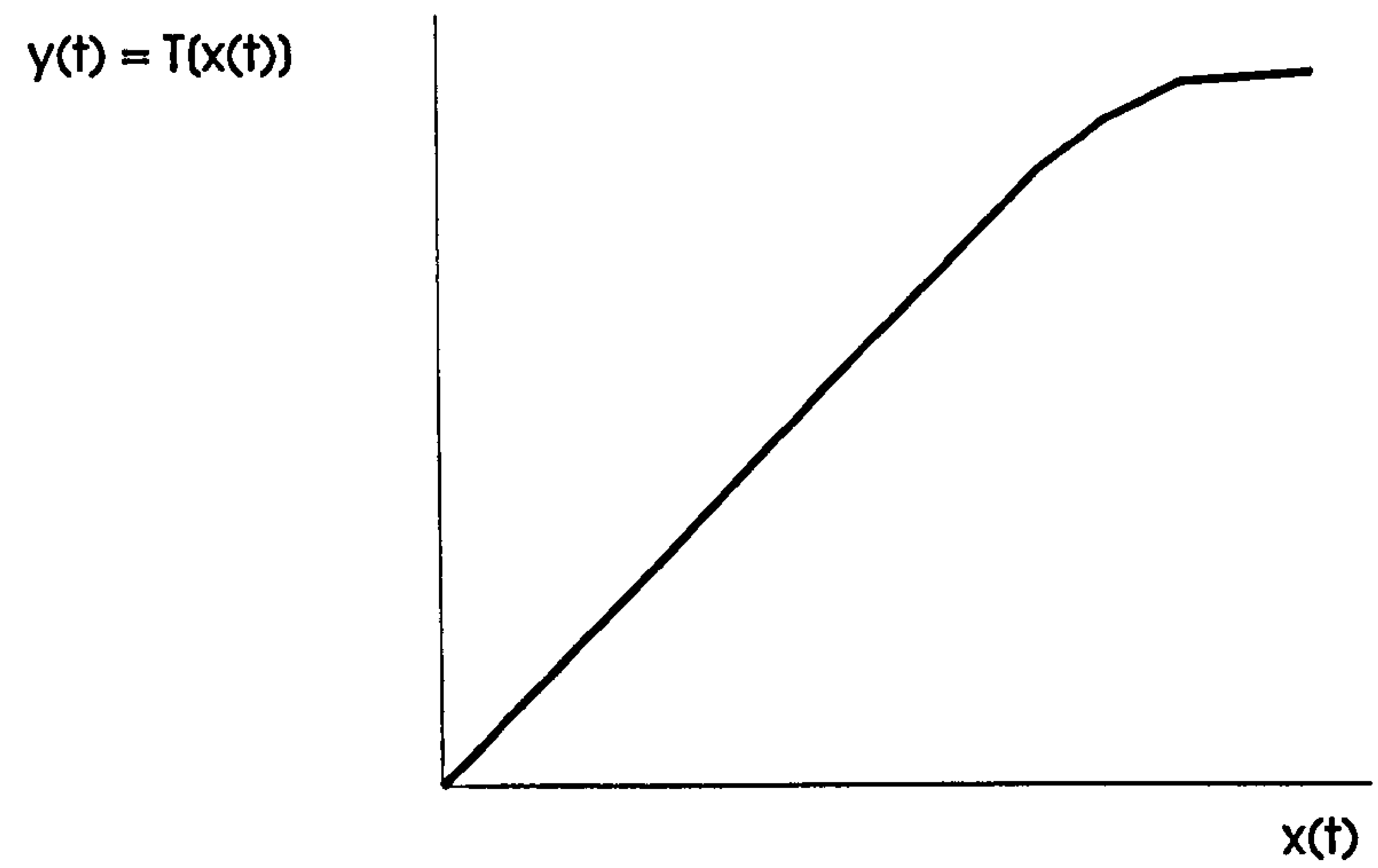

Figure 2.1 Transfer Characteristic of a Non-linear System

2.2.2 Amplitude Distortion in Power Amplifiers

The most common type of non-linearity used to characterise power amplifiers is the amplitude, or more commonly the AM-AM distortion introduced by the amplifier being considered. If the amplifier is assumed to be memory less, that is, its output voltage is an instantaneous function of its input voltage. Then the output voltage $e_0(t)$ can be represented by a power series [1] of the input voltage $e_i(t)$ as

$$e_0 = k_1 e_i + k_2 e_i^2 + k_3 e_i^3 + \ldots + k_n e_i^n \tag{2.3}$$

For a linear amplifier all k_i's are zero for $i = 2,3,\ldots,$. The representation of e_0 in equation 2.3 neglects the phase characteristic of the amplifier. Considering an amplifier with a mild non-linearity such as a typical class A or AB amplifier then e_0 may be represented as

$$e_0 = k_1 e_i + k_2 e_i^2 + k_3 e_i^3 \tag{2.4}$$

If $e_i = A\cos\omega_1 t$ then e_0 can be written as

$$e_0 = \frac{1}{2}k_2 A^2 + \left(k_1 A + \frac{3}{4}k_3 A^3\right)\cos\omega_1 t + \frac{1}{2}k_2 A^2 \cos 2\omega_1 t + \frac{1}{4}k_3 A^3 \cos 3\omega_1 t \tag{2.5}$$

Equation 2.5 shows how the resultant signal consists of the fundamental frequency ω_1 and additional signals at dc, the second harmonic frequency $2\omega_1$, and the third harmonic frequency $3\omega_1$. In a practical system these out of band components may be filtered out and so they will have no impact on bandwidth performance of the operational system. From equation 2.5 it can be seen that the fundamental component of e_0 has an amplitude of $k_1 A\left(1+\frac{3}{4}\left(\frac{k_3}{k_1}\right)A^2\right)$, which is greater than $k_1 A$ (the linear gain) if $K_3 > 0$ and smaller than $k_1 A$ if $K_3 < 0$. This is known as gain expansion and gain compression respectively. Practical amplifiers are compressive i.e. $K_3 < 0$ and their output power is usually characterised at the 1dB compression point. The 1dB compression point being defined as the output level at which the amplifiers gain is 1dB less than the small signal gain. An amplifier may be characterised in this way by applying a range of input powers to the amplifier and measuring its gain.

Intermodulation Distortion and Intercept Point

Changes of gain with amplitude cause the output signal to be distorted. If a two tone signal is applied to this system such that

$$e_i = A(\cos\omega_1 t + \cos\omega_2 t) \tag{2.6}$$

it may be shown [1]

$$\begin{aligned} e_0 = {} & k_2 A^2 + k_2 A^2 \cos(\omega_1 - \omega_2)t + \left(k_1 A + \frac{9}{4}k_3 A^3\right)\cos\omega_1 t + \left(k_1 A + \frac{9}{4}k_3 A^3\right)\cos\omega_2 t \\ & + \frac{3}{4}k_3 A^3 \cos(2\omega_1 - \omega_2)t + \frac{3}{4}k_3 A^3 \cos(2\omega_2 - \omega_1)t + k_2 A^2 \cos(\omega_1 + \omega_2)t \\ & + \frac{1}{2}k_2 A^2 \cos 2\omega_1 t + \frac{1}{2}k_2 A^2 \cos 2\omega_2 t + \frac{2}{4}k_3 A^3 \cos(2\omega_1 + \omega_2)t + \frac{3}{4}k_3 A^3 \cos(2\omega_2 + \omega_1)t \\ & + \frac{1}{4}k_3 A^3 \cos 3\omega_1 t + \frac{1}{4}k_3 A^3 \cos 3\omega_2 t \end{aligned} \tag{2.7}$$

Equation 2.7 shows that the output signal consists of dc terms, the fundamental terms ω_1 and ω_2, second harmonic terms at $2\omega_1$ and $2\omega_2$ and third harmonic terms at $3\omega_1$ and $3\omega_2$. The undesired harmonic terms may be filtered to leave only the fundamental tones. In addition intermodulation products are also produced. These consist of the second order intermodulation products (IMP's) at $\omega_1 \pm \omega_2$ and the third order intermodulation products at $2\omega_1 \pm \omega_2$ and $2\omega_2 \pm \omega_1$. The second order IMP's fall outside the passband and therefore are filtered in the same way as the dc terms and the harmonics. The third order IMP's however fall within the passband and will distort the desired output signal at the fundamental frequencies ω_1 and ω_2, this is known as intermodulation distortion (IMD)[1]. This is shown in figure 2.2.

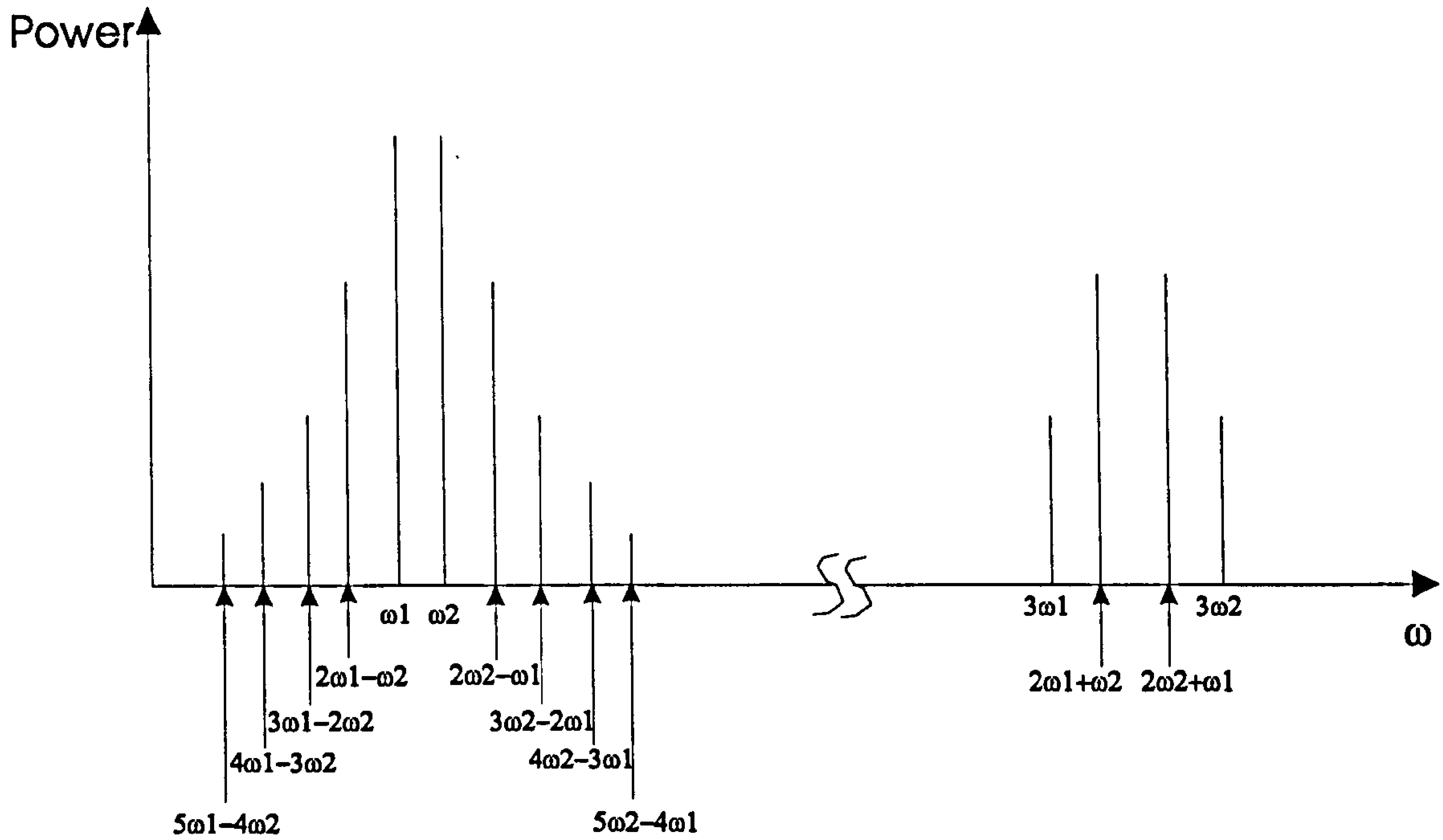

Figure 2.2 Spectrum of Intermodulation Distortion

This result has important implications when assessing amplifier linearity, the higher the IMP power the poorer the linearity. A two-tone test is used as the main test of the linearity performance of an amplifier being linearised, or an amplifier being used in an operational system. The lower the third order and subsequent orders of IMP's are below the main tones the better the performance of the amplifier[2]. The two-tone test is usually carried out at the 1dB compression point, this gives a repeatable point for comparison of results between different types of amplifiers and linearisation schemes. The reasons for choosing this point are twofold, firstly the 1dB compression point is an easily measured amplifier parameter and secondly the third order IMP's increase at three times the rate of the main tones when the amplifier is driven into compression, therefore for every 1dB increase in the main tones the third order IMP's increase by 3dB. This leads to the concept of the third order intercept point.

This is defined as the output power level P_I at which the output power $P_{(2\omega_1-\omega_1)}$ at the frequency $2\omega_1 - \omega_2$ would intercept the output power P_O at ω_1 (when the amplifier is

[1] This measurement is usually carried out at the 1dB compression point.
[2] This is usually measured in dBc.

linear) if low-level results were extrapolated into the higher power region [1] as shown in figure 2.3.

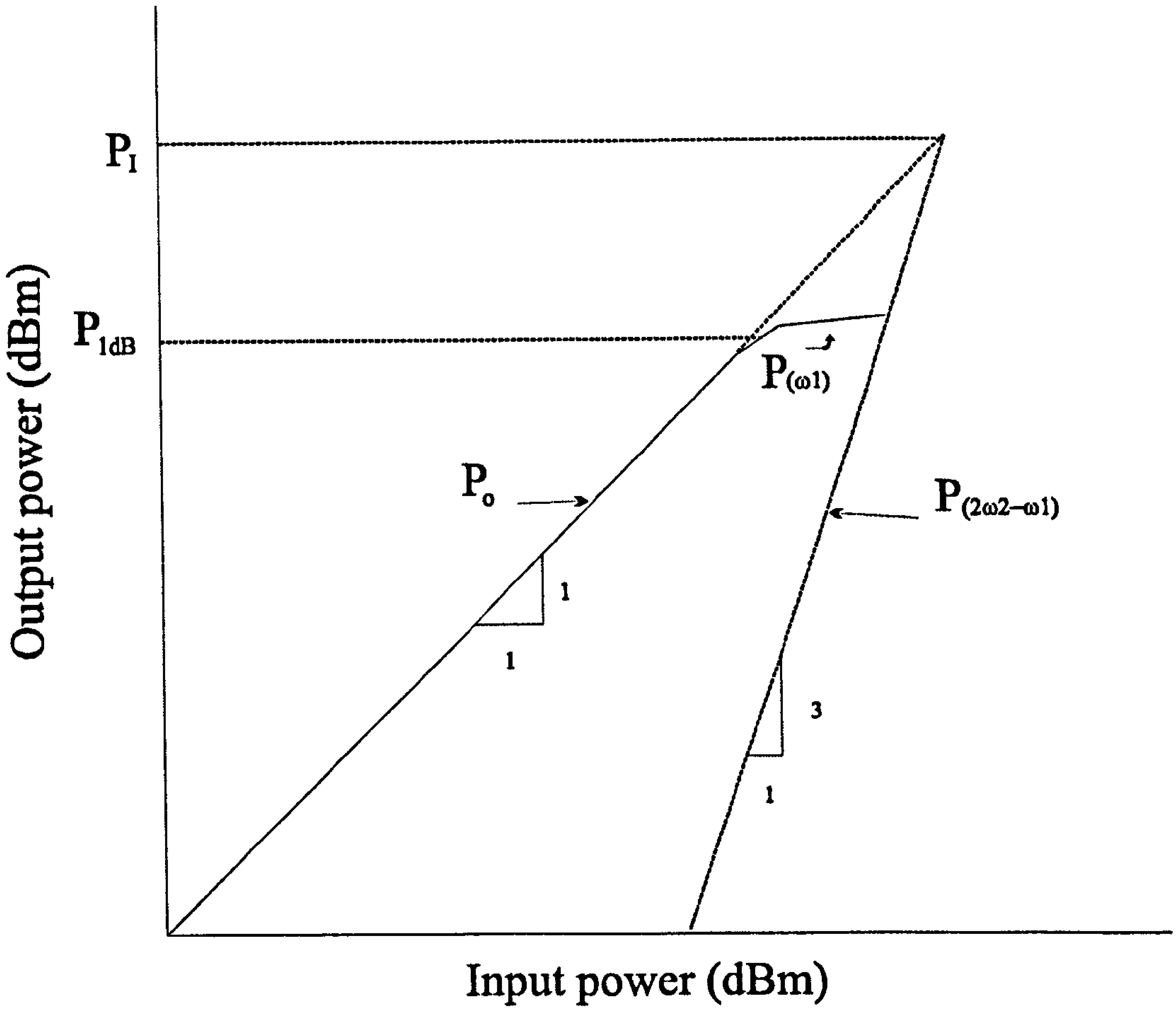

Figure 2.3 Definition of Intercept Point

Figure 2.3 shows that at low power levels the output power P_O is directly proportional to the amplitude of the input signal while the output power $2\omega_2 - \omega_1$ is directly proportional to the cube of the input amplitude. Therefore on a log-log scale (i.e. dBm-dBm scale) the plot of each will be a straight line with a slope corresponding to the order of the response, i.e. the response at ω_1 will have a slope of 1 and the response at $2\omega_1 - \omega_2$ will have a slope of 3, where they intersect is the intercept point. It may be noted that the intercept power P_I is independent of the input power and is therefore a useful measure of the system non-linearity, which is quoted in manufacturers literature when comparing amplifiers.

2.2.3 Phase Distortion in Power Amplifiers

Apart from amplitude distortion amplifiers can be affected by another type of distortion, where the phase shift is a function of the instantaneous amplitude of the input signal. The output phase can have a ripple around a mean phase value. This is known as AM to PM conversion. The amount of AM to PM associated with a particular amplifier can be measured by applying various input powers to the amplifier and measuring the phase change across the amplifier. The AM to PM transfer characteristic can then be produced which relates power level to phase change.

2.3 Bandwidth of Operation

The bandwidth over which amplifiers operate may be described as being either wideband or narrowband. In general these terms are note well defined, and so a more satisfactory definition is required[3].

Relative bandwidth is expressed as a percentage, and is defined as

$$B_{rel} = \frac{B}{f_o} \tag{2.8}$$

where B is the absolute bandwidth, and f_o is the centre frequency of the band. This is an important parameter in the characterisation of amplifiers and auxiliary components.

Channel bandwidth is the absolute bandwidth, which a single channel occupies in a multi-channel system.

[3] The author acknowledges Dr. A. Mansell as the source of this definition

For the purposes of this work, the definitions in Table 2.1 are used.

Term	Relative Bandwidth	Channel Bandwidth
Narrowband	< 0.5%	< 30kHz
Mediumband	0.5 – 5.0%	30 – 500kHz
Wideband	> 5.0%	> 500kHz

Table 2.1 Definitions of Bandwidth

2.4 Traditional Amplifier Classes

The previous sections have shown how amplifiers maybe characterised and compared in terms of their performance. This section will give a brief overview of the traditional amplification classes [2].

2.4.1 Class A

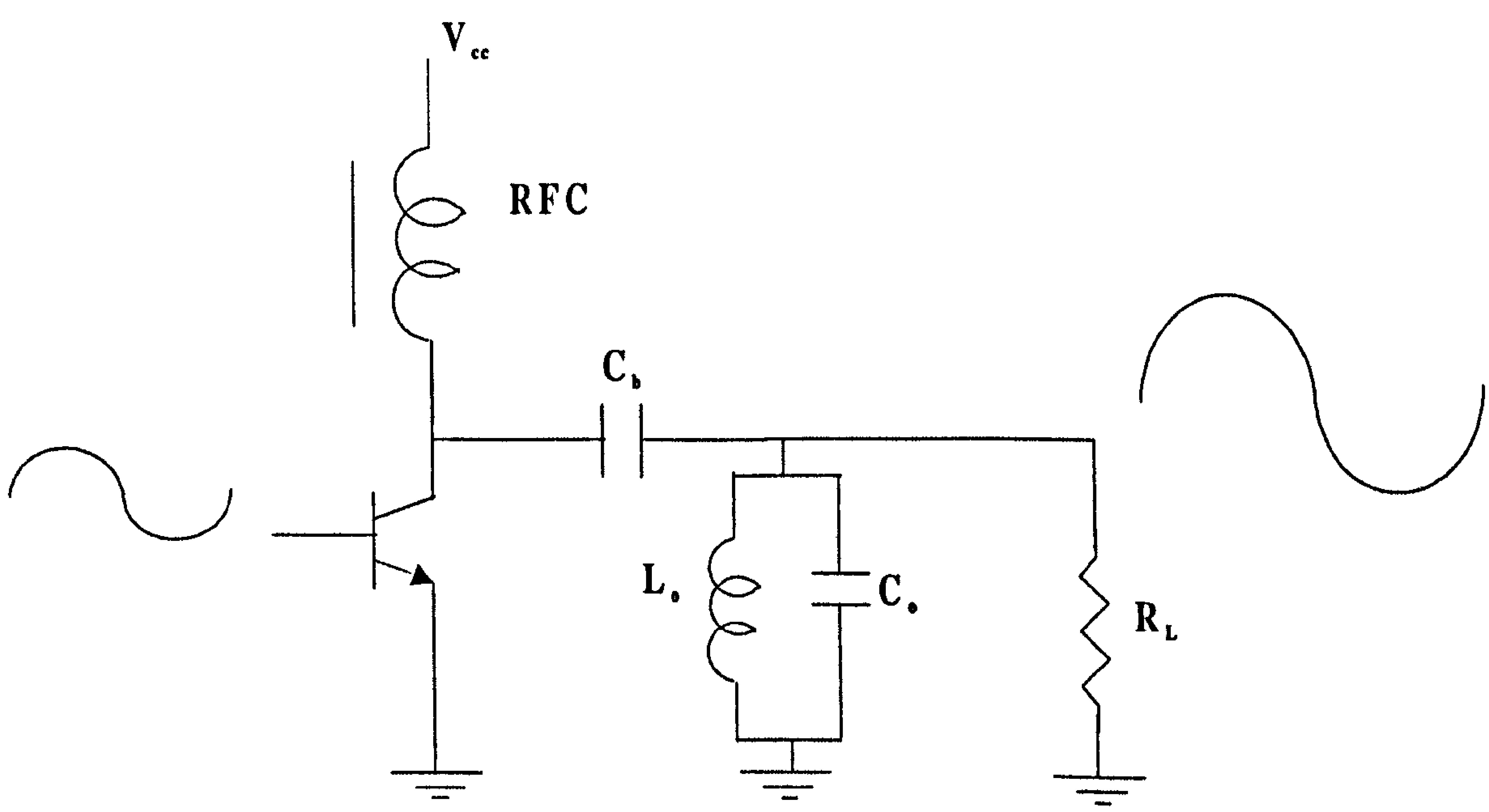

Figure 2.4 Class A R.F. Power Amplifier

Class A power amplifiers operate with a continually flowing bias current which is arranged such that the transistor is operating within it's linear region of operation. A typical class A power amplifier is shown in figure 2.4. This makes class A power amplifiers the most linear of the traditional classes of amplifier. Typical figures for linearity for a class A amplifier are between –30 to –60dBc. This large variation in linearity performance is due to the method of operation of the amplifier. It is possible to improve the linearity performance of the amplifier by increasing the amount of back off applied to the amplifier. This has an impact on the achievable efficiency of a class A amplifier. The efficiency of a class A amplifier maybe expressed as [2]:

$$\eta = \frac{P_o}{P_i} = \frac{V_o^2}{2V_{CC}^2} \leq \frac{1}{2} \tag{2.9}$$

This is the maximum theoretical efficiency assuming that the amplifier will be driven into saturation. When linearity is important then significant amounts of back off will be required which has a severe impact on the efficiency of the amplifier so in practice measured efficiencies are lower than the theoretical maximum, typical values are 10-15%.

2.4.2 Class B

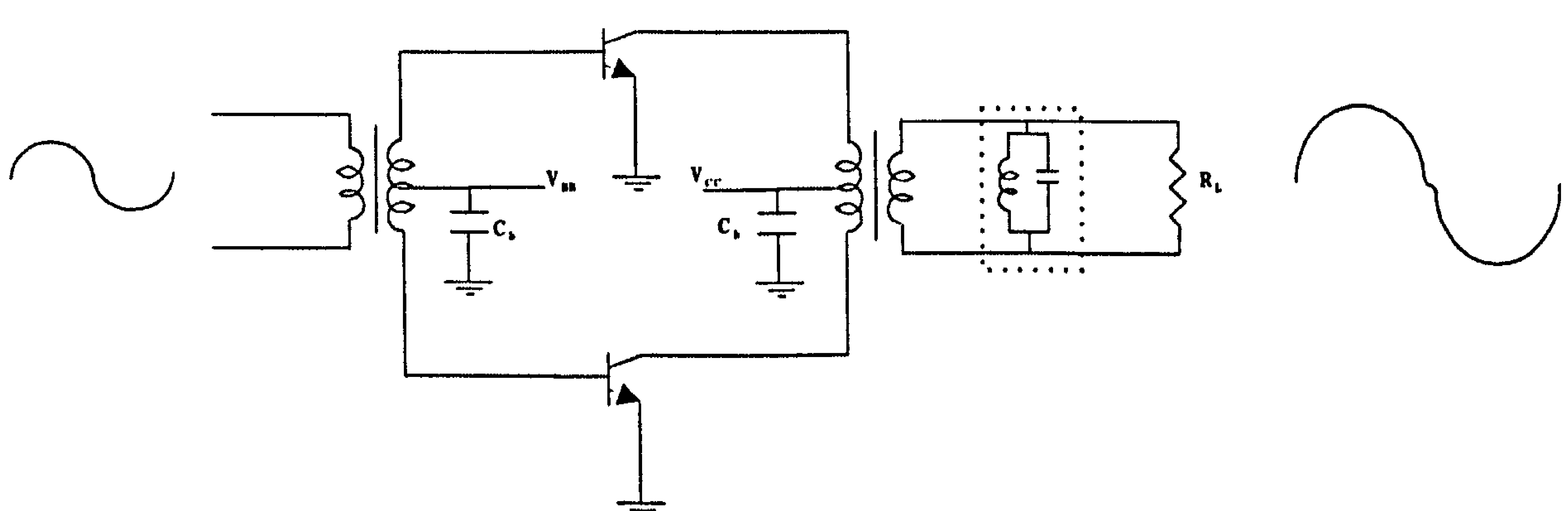

Figure 2.5 Class B R.F. Power Amplifier

Class B amplifiers employ two transistors in an effort to improve the poor efficiency of class A power amplifiers. A typical class B power amplifier is illustrated in figure 2.5. The amplifier operates in the following manner, the input signal is split into positive and negative half cycles these signals are then amplified by power transistors and then recombined at the output to generate an amplified composite signal. The transistors are biased at cut off so that no current flows until a signal is applied. This has the effect of increasing the efficiency of this type amplifier with respect to a class A amplifier at the expense of increasing complexity. The efficiency of a class B may be calculated from [2]:

$$\eta = \frac{P_o}{P_i} = \frac{\pi V_o}{4V_{CC}} \leq \frac{\pi}{4} \approx 78.5\% \tag{2.10}$$

This equation shows that the efficiency of a class B amplifier is considerably higher than a class A amplifier. But this improvement in efficiency is obtained at the expense of amplifier linearity. Typical practical conversion efficiency is usually 30-50%. Typical values of distortion for this amplifier are –20to -25dBc. The reduced linearity performance of this type of amplifier is due to operating the amplifier just at cut off, this introduces distortion which is caused by the transistor turning on, this distortion effect is known as cross over distortion

2.4.3 Class AB

Class AB amplifiers are class B amplifiers which are biased such that a small amount of current follows even when no signal is present. This current is designed to eliminate to a large extent the distortion due to the cross over effect. This approach does have the disadvantage of reducing amplifier efficiency to some extent. But if the bias current is chosen to be an extremely small proportion of the signal current then the efficiencies obtained will still be greater than those obtainable with a class A amplifier, while an amplifier with a linearity performance approaching that of a class A amplifier is produced.

2.4.4 Class C

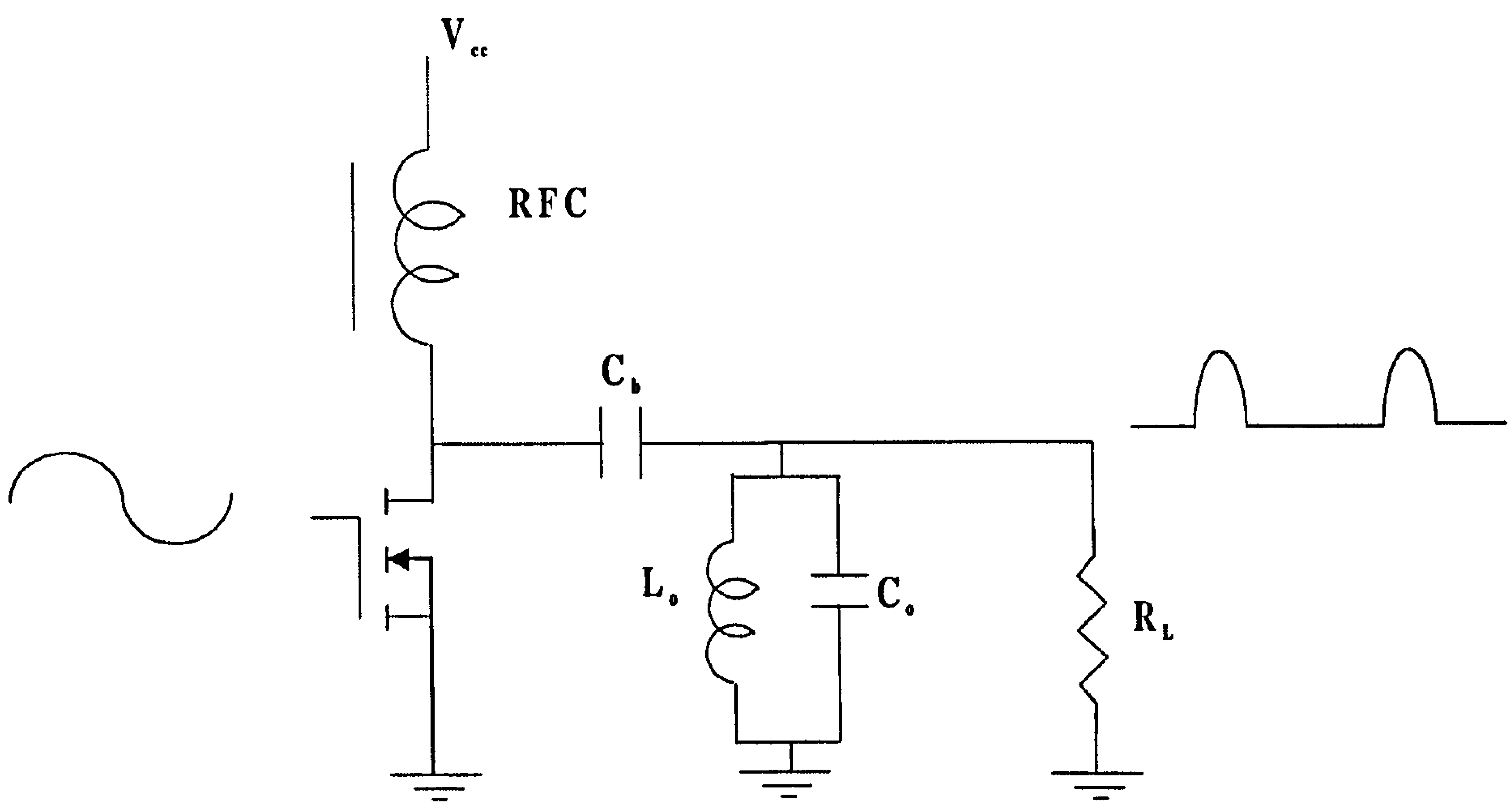

Figure 2.6 Class C R.F. Power Amplifier

Class C power amplifiers operate by biasing the transistor strongly into the cut-off region. A class C amplifier is shown in figure 2.6. This means that the conduction angle is much less than 180^{0}, which results in an amplifier which has the highest efficiency of any of the non-switched amplifier classes. The efficiency of this amplifier is related to the conduction angle that is chosen by the designer. Class C amplifiers are usually used as saturated amplifiers for at least part of the conduction cycle to increase the overall efficiency of the amplifier. This enables efficiencies of the order of 50-60% to be achieved in practical applications. Early class C amplifiers were designed to operate with vacuum tubes [3], but the biasing techniques used for vacuum tubes may not be used for solid state power amplifier applications. To bias a solid state Class C amplifier a technique known as "Class C mixed mode" is used [2]. Due to the extremes of operation of class C amplifiers their intermodulation performance is poor with the highest products being typically at –15dBc.

2.4.5 Classes D, E and S

Switching amplifiers are amplifiers in which the transistors are employed as switches rather than as current sources. This reduces the power dissipation within the device and thus increases the efficiency that may be achieved. Switching amplifiers may be used within broadcast transmitter applications and are finding application within certain linearisation schemes such as LINC, this and other linearisation techniques are reviewed later in this chapter.

Class D amplifiers [4] employ two active devices that are operated as two pole switches. These devices are connected to a tuned output circuit that removes the harmonics resulting in a sinusoidal output. Class D amplifiers can be used as broadband amplifiers, but their high frequency performance is poor due to device capacitance. The efficiency of an ideal class D amplifier is 100%.

Class E amplifiers [5] use a single device connected to a passive load network. The amplifier utilises a shunt capacitance to reduce the effect of device capacitance so the overall high frequency performance of this amplifier is superior to class D amplification. Class E amplifiers can be used in high frequency applications but due to the tuned nature of the shunt capacitance the amplifiers may only be used in narrowband applications. The theoretical efficiency of a class E amplifier is 100% but practical efficiencies are usually 80%.

Class S amplifiers [6] employ a PWM technique to sample the signal at many times the operating frequency. The need to sample the waveform limits the maximum frequency of operation of this type of amplifier. To obtain a meaningful signal usually requires a sample rate of 2.5-3 times the operating frequency. Again the theoretical efficiency of this type of amplifier is 100%, but practical amplifiers are usually around 70-80% efficient due to device imperfections.

2.5 Linearisation Techniques

Linearisation techniques can be divided into two broad categories, those that are narrowband and those that are broadband. The linearity improvement that is obtained is dependent on the technique employed and on the amount of non-linearity already present in the amplifier being linearised. This section describes some of these techniques and explains their advantages and limitations.

2.5.1 Narrowband Techniques

2.5.1.1 R. F. Feedback

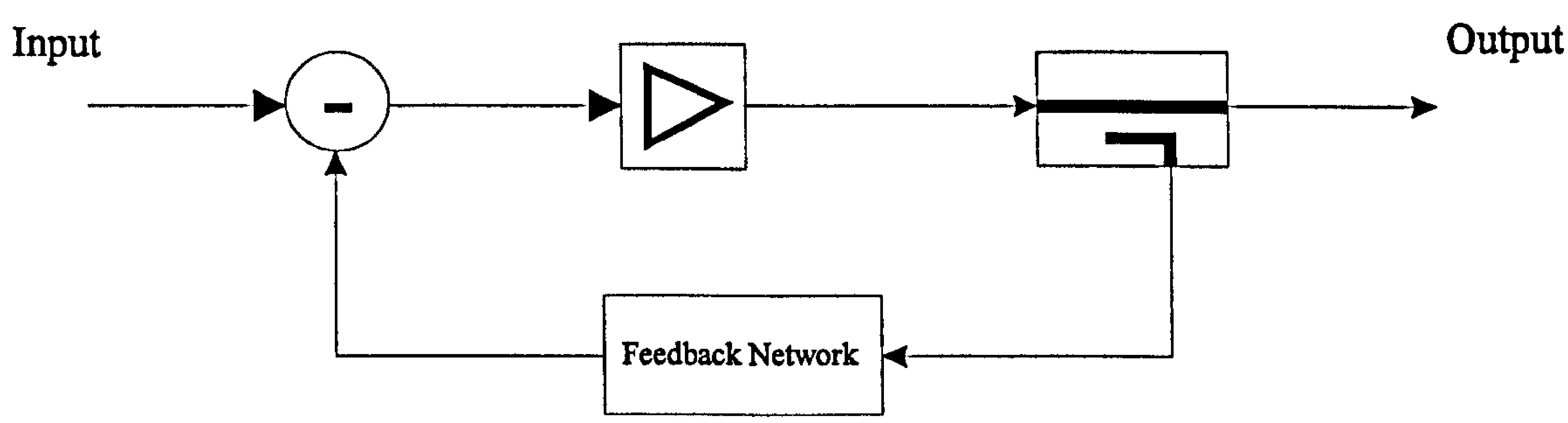

Figure 2.7 R. F. Feedback

R. F. feedback [7] uses a high gain amplifier whose gain is reduced by feedback so that it operates in a more linear manner, a typical R.F. feedback amplifier is shown in figure 2.7. This method is potentially unstable due to the use of feedback. Under normal operation the signal is fed-back 180 degrees out of phase to the input resulting in linear operation. But if a further 180 degrees of phase shift is introduced through the loop, then positive feedback results and oscillation will occur. R. F. Feedback is a narrowband technique due to the feedback process.

2.5.1.2 Cartesian Loop

Cartesian loop [8, 9, 10] shown in figure 2.8 is primarily a feedback technique.

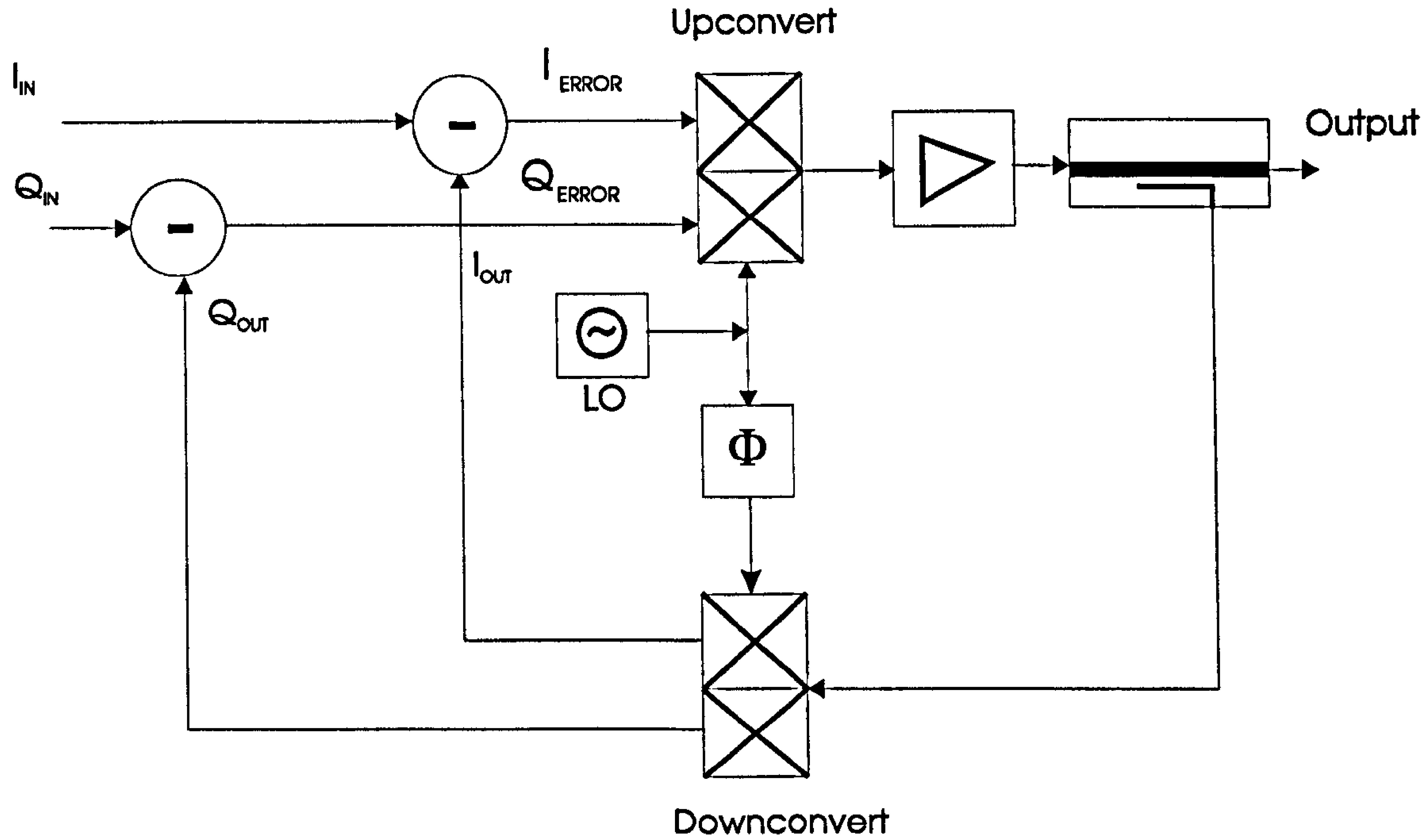

Figure 2.8 The Cartesian Loop Amplifier

The signal consists of quadrature components, the I & Q signals. These signals are fed into differential amplifiers that implement the subtraction process to generate the error signals. The outputs of the differential amplifiers are up-converted to R.F. using a quadrature R.F. oscillator. The R.F. signals from the two paths are then combined to the R.F. output signal that is then applied to the amplifier.

Cartesian loop does have several draw backs however. To achieve high linearity performance the loop must be stable, the cartesian signals must be accurately generated and the quadrature of the local oscillators must be maintained. Loop stability may be improved by inserting a phase shifter between the LO inputs of the quadrature up and down conversion paths which increases the phase margin of the loops. The required phase change is a function of frequency so the stability can only be maintained for narrowbandwidths.

This method does provide excellent linearity over narrowbandwidths and amplifiers with up to -75dBc performance have been designed at a variety of frequency bands in VHF, UHF and SHF.

2.5.1.3 Polar Loop

Polar loop is in effect a technique [11] very similar to cartesian loop, but instead of using the I and Q baseband cartesian components it utilises the baseband amplitude and phase components i. e. the polar components, the technique is illustrated in figure 2.9.

Polar loop in effect utilises a phase locked loop to up and down convert the baseband signals. The baseband single sideband signal is used to control the operating frequency of the VCO which controls the frequency of operation of the transmitter. The original baseband SSB signal and the down converted transmitter output are compared and used to provide the feedback action which is applied to the RF power amplifier power supply. This form of transmitter has to date mainly been used for narrowband linearisation due to the problems of feedback stability. The results for narrowband operation are reasonable with IMD products suppressed to less than -50dBc with tone spacing of 0.5kHz.

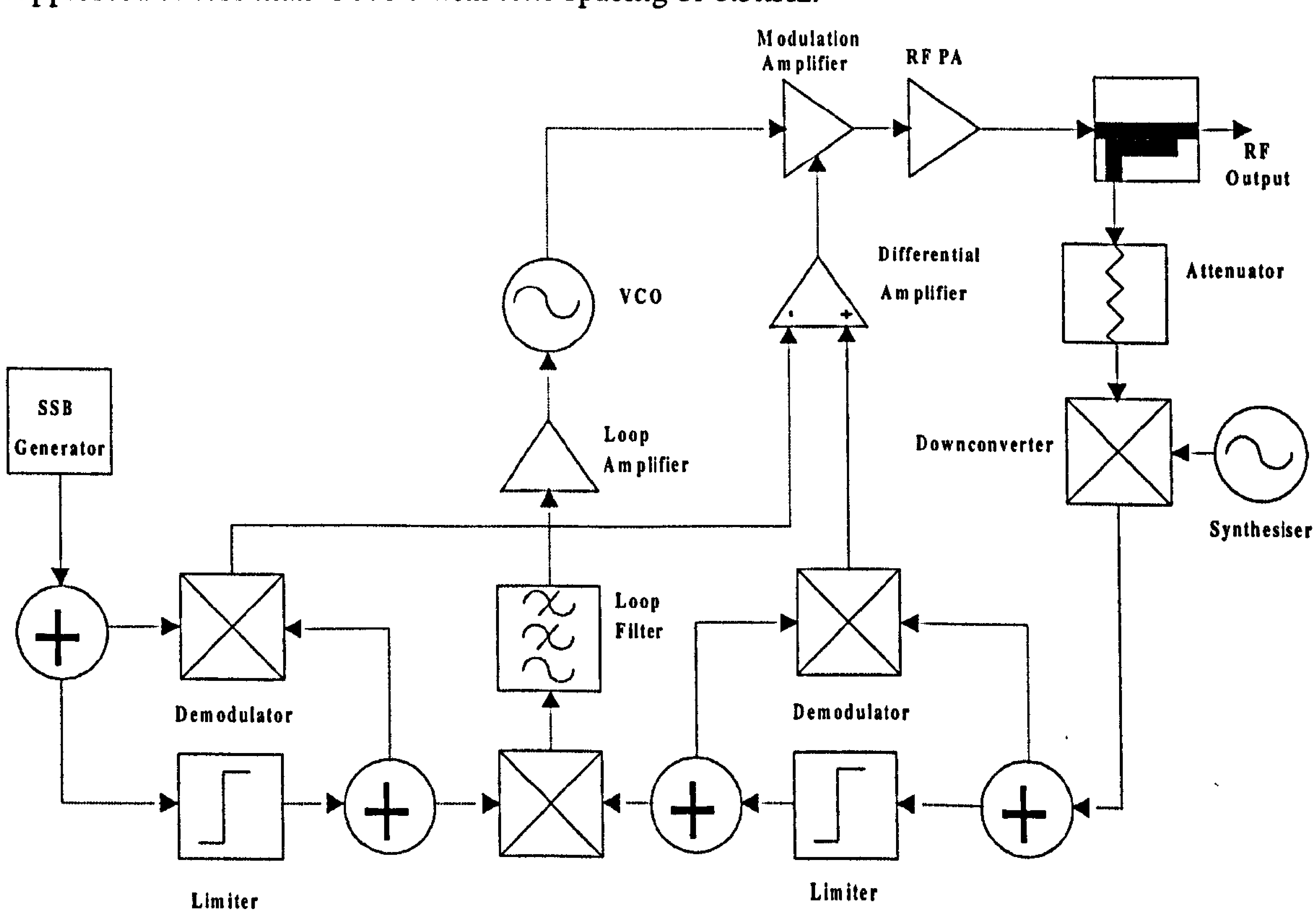

Figure 2.9 Polar Loop Linearisation Scheme

2.5.1.4 Adaptive Baseband Predistortion

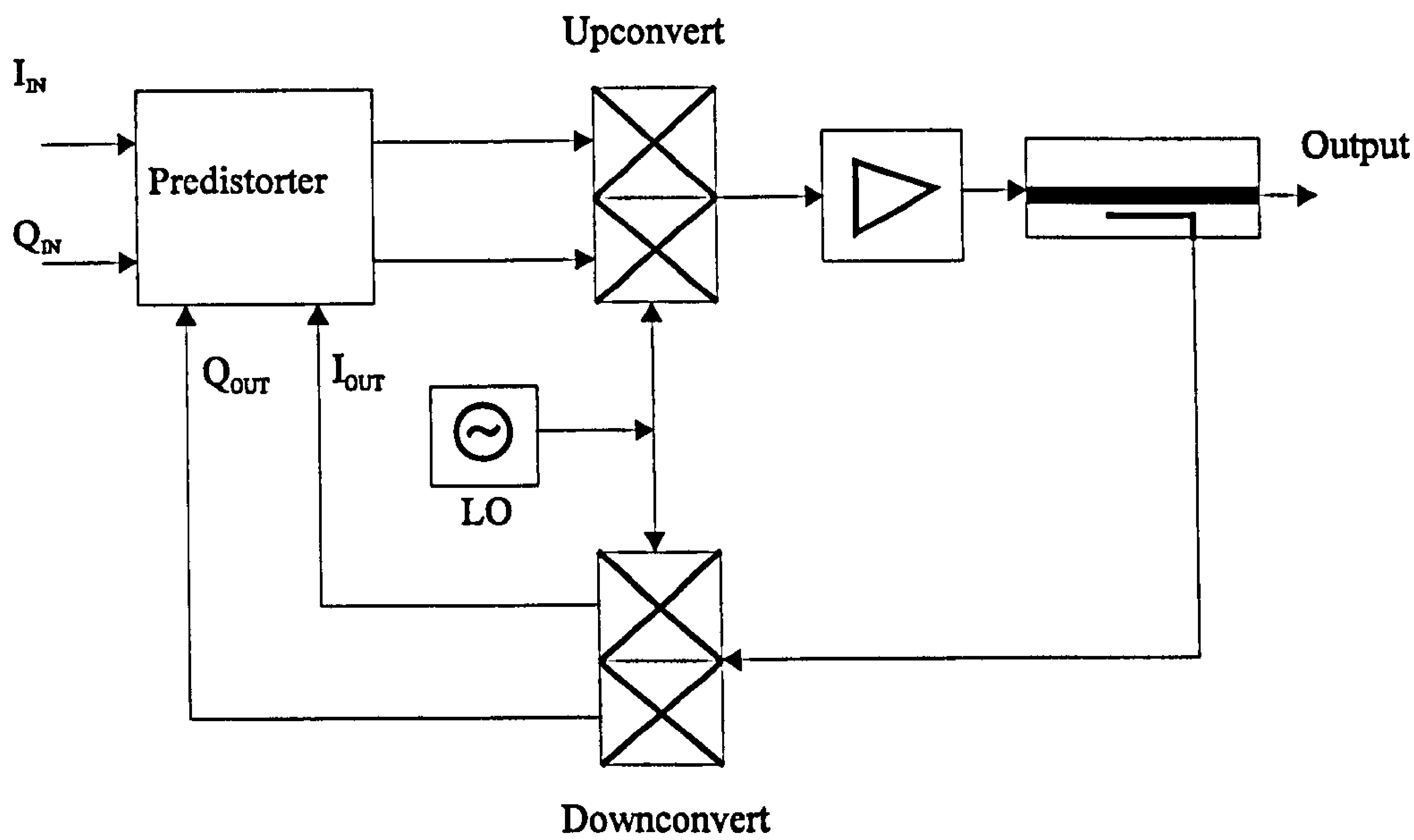

Figure 2.10 Adaptive Baseband Predistortion

Adaptive baseband predistortion shown in figure 2.10, uses a DSP to calculate the coefficients of the predistorter required to produce a linear output. There are three basic forms; mapping based, gain based and analogue, these are discussed below.

Mapping Based Predistorter

A mapping based predistorter uses a two dimensional look up table which maps any point on the complex plain to any other point. The input signal in I & Q format is converted by the DSP into a predistorted signal by direct mapping. This signal is then up-converted and amplified. A part of the output signal is then down-converted and fed-back into the predistorter and used to adapt the look up table to minimise the error between each I & Q constellation point. This has the effect of minimising the distortion. This method has the potential to provide excellent linearity with IMP's of -60dBc achieved through simulation.

The technique does have several disadvantages however. To obtain good cancellation the look up table for 10 bit accuracy needs to have 20Mbits of data. The time of convergence is slow typically 10 seconds. Due to the use of a DSP the technique is inherently narrow-band. Finally the power consumption of the DSP and associated ADC's and DAC's will be a significant proportion of the total power requirement of a typical hand portable unit. These devices may consume up to 4W [12] which is significantly higher than the amplifier alone which normally consumes less than 1W in most systems and less than 0.5W in many systems. This means that the power efficiency of an adaptive baseband system is very poor. This will continue to be the case until device technology advances to the point where the predistorter power consumption is a small proportion of the amplifier power consumption.

Gain Based Predistortion

This method of predistortion is effectively a one dimensional [13] mapping predistorter. It uses the envelope level of the signal to produce the required complex output signals. Interpolation is used to calculate the required intermediate values. Because this method only refers to the amplitude to calculate it's coefficient's the memory requirement is much lower, typically only approximately 64 words are required to produce similar performance to the mapped based technique. The convergence time is much lower due to the reduction in the number of mapped points (typically milliseconds this is dependent on the channel bandwidth), and the re-convergence time is negligible.

This method however is much more computationally intensive than the mapped based technique. This is due to the necessity of calculating the interpolated points. Additionally accurate control of the cartesian modulator is required, as the method cannot correct errors due to the modulation process.

Baseband Analogue Predistorter

This method [14] uses analogue components to predistort the input signal which are controlled by a DSP. The predistorter characteristic is not altered in real time and so convergence time is relatively slow. The amplifier output is down-converted to baseband and the envelope used by the DSP to calculate the new predistorter characteristics. This technique is much less complex than the other adaptive baseband predistortion techniques.

2.5.1.5 RF Synthesis

RF Synthesis techniques are a group of amplifier linearisation schemes where the RF signal is split up into separate constant envelope components which are amplified using high efficiency non-linear amplifiers and then recombined to form a composite linearly amplified output signal.

Envelope Elimination and Restoration (EE & R)

Envelope elimination and restoration was first proposed by Kahn in 1952 [15] and was further developed for use as a single sideband transmitter [16]. The technique is essentially an RF amplifier linearisation technique. The EE & R technique is now extensively used in high power TV and radio transmitters. A block diagram of the technique is shown in figure 2.11.

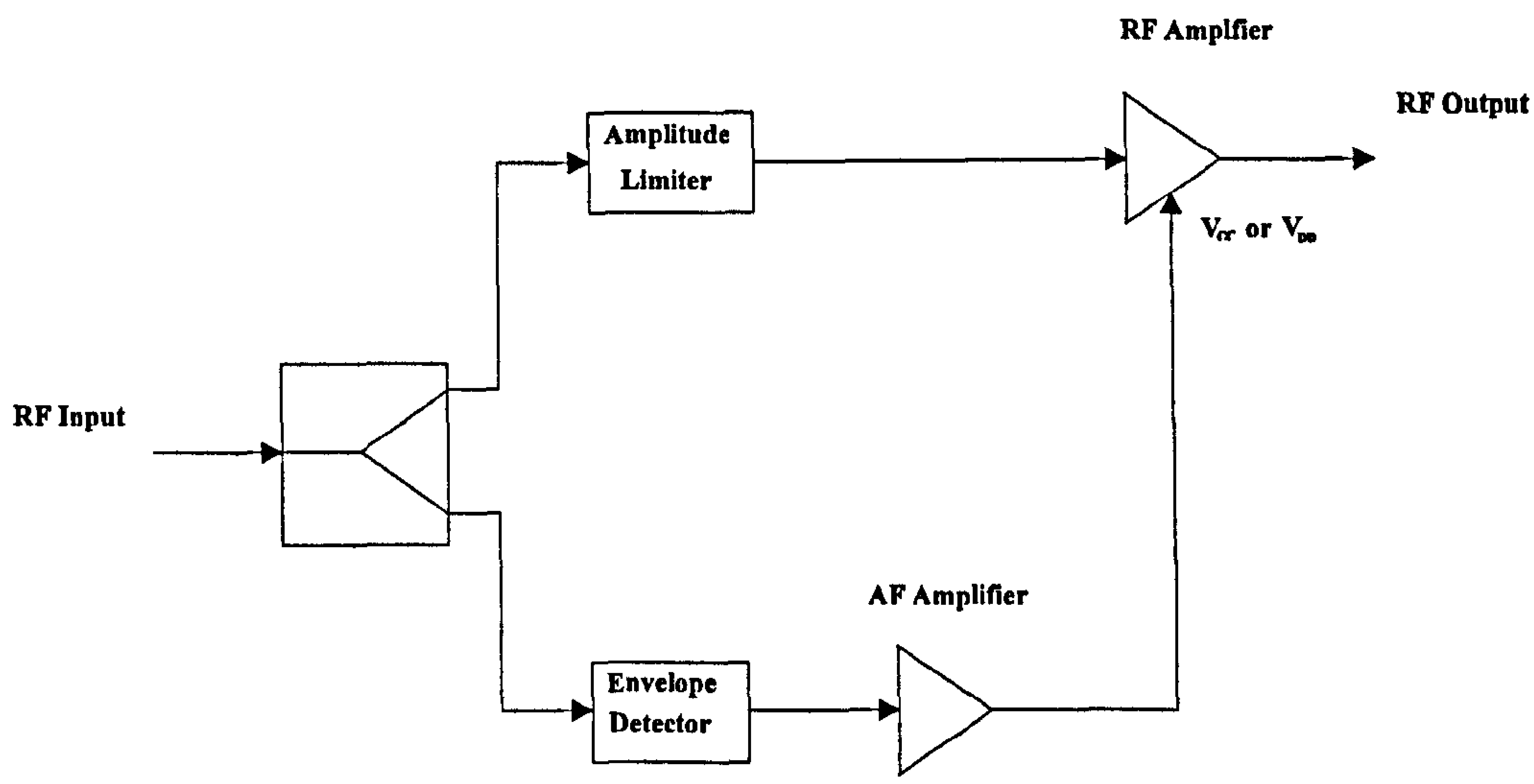

Figure 2.11 The Envelope Elimination and Restoration Amplifier

The technique works by splitting the input signal into two separate signals, an amplitude modulated signal and a phase modulated signal. The phase modulated signal is amplified by an efficient R.F. amplifier, usually a switching amplifier is used at VHF frequencies and class C amplifiers are preferred for UHF frequencies. The amplitude modulated signal is amplified by an efficient A.F. amplifier, the output from this amplifier being used to modulate the R.F. amplifiers power supply.

LInear Amplification using Non-Linear Components (LINC)

LINC originally suggested by Cox [17] in 1974 uses the principle of splitting the composite baseband signal into separate constant amplitude phase varying signals. These signals may be amplified separately by any amplifier with sufficient bandwidth for the task no matter how non-linear that amplifier may be. This technique is illustrated in figure 2.12.

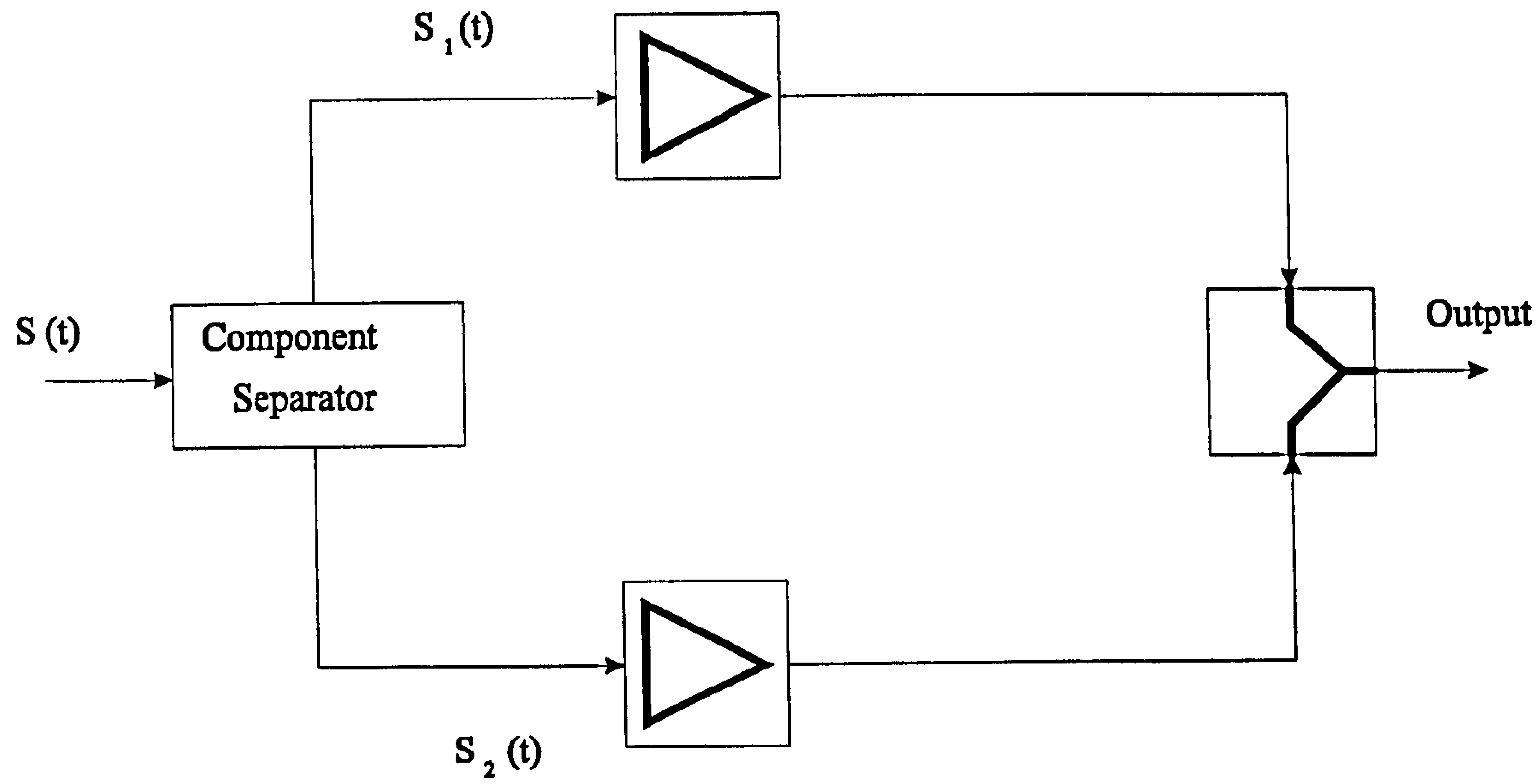

Figure 2.12 The LINC Transmitter

Mathematically this may be expressed as

$$S(t) = E(t)\cos(\omega_c t + \theta(t)) \tag{2.11}$$

$$S_1(t) = E_m \cos(\omega_c t + \theta(t) + \alpha(t)) \tag{2.12}$$

$$S_2(t)E_m \cos(\omega_c(t) + \theta(t) - \alpha(t)) \tag{2.13}$$

$$\alpha(t) = \cos^{-1}\left(\frac{E(t)}{E_m}\right) \tag{2.14}$$

where: $S(t)$ is an arbitrarily modulated signal

$E(t)$ is the amplitude of the signal which has some maximum value E_m

ω_c is the carrier frequency

$\theta(t)$ is the phase content of the signal

$S_1(t)$ and $S_2(t)$ are the resultant constant envelope signals

The LINC technique has the potential to provide almost 100% efficient amplification if 100% efficient amplifiers were used. The technique does however have a number of problems if an effective implementation is to be achieved, first the signals have to be split in a very precise manner and the bandwidth occupied by these signals prevents the technique being used for broadband applications. The second problem is the necessity for a low loss high power combiner for the resultant amplified amplitude and phase signals, some work has been undertaken in this area [18] but still this issue is likely to prevent the LINC technique from achieving it's promised levels of efficiency.

The Combined Analogue Locked Loop Universal Modulator (CALLUM)

The CALLUM [19, 20, 21] technique is effectively a LINC derivative which addresses the component separation, system control and frequency translation issues necessary for an effective transmitter of this type. The CALLUM transmitter is shown in figure 2.13, the system incorporates a feedback control loop which carries out all the necessary signal conditioning and control for the effective operation of the transmitter. The main difference between this transmitter and the standard LINC transmitter is that the control and frequency translation is carried out using a VCO for each path. CALLUM may be implemented in analogue or DSP forms. In it's analogue form it is a definite contender for use as the mobile part of the TETRA radio system. CALLUM however due to it's feedback topology is only applicable to narrowband systems.

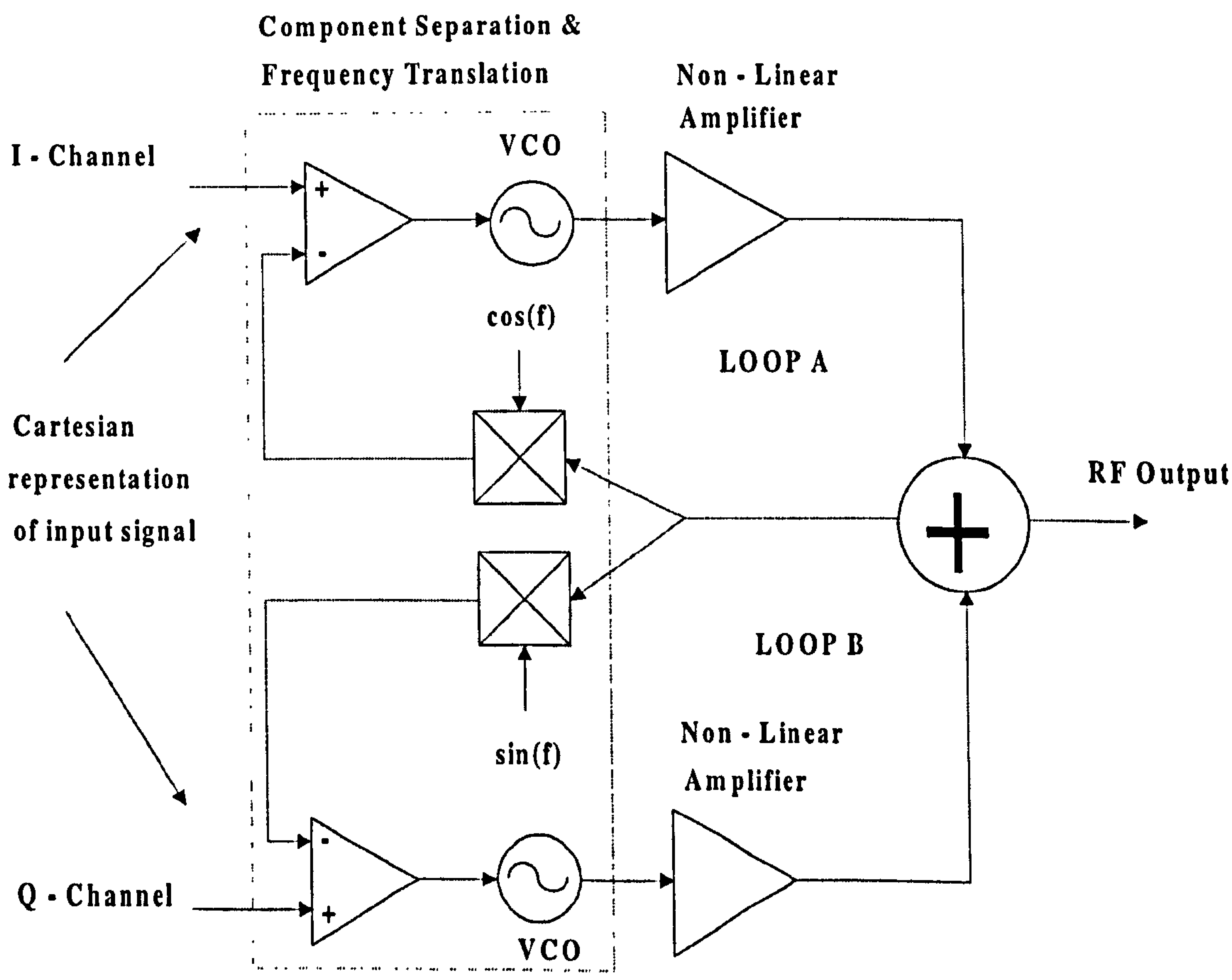

Figure 2.13 The Combined Analogue Locked Loop Universal Modulator

Linear Amplification by Sampling Techniques (LIST)

The LIST technique was originally proposed by Cox in 1975 [22], the technique is essentially a modified form of the LINC transmitter which uses digital sampling techniques to carry out the signal separation. A diagram of a typical LIST transmitter is shown in figure 2.14. The principle is to separate the signals into two constant amplitude signals that have only discrete reversals in phase. These signals are than amplified separately by any amplifier or phase locked oscillator that has the required bandwidth regardless of the amplitude linearity. The LIST technique still suffers form the draw back of the LINC techniques in general i.e. the difficulties of sampling and generating the two constant amplitude waveforms and the difficulties of efficiently power combining the two independently amplified signals. A serious drawback of the technique is the high sample rate required by the technique for reasonable signal to noise ratios, for example for a multi-carrier system with a bandwidth of 20MHz with a SNR of 30dB a sample rate of 450MHz is required. The technique is quoted in practical tests of being capable of 40dB of IMP suppression for a 50kHz-bandwidth signal, this limits the technique to narrowband applications.

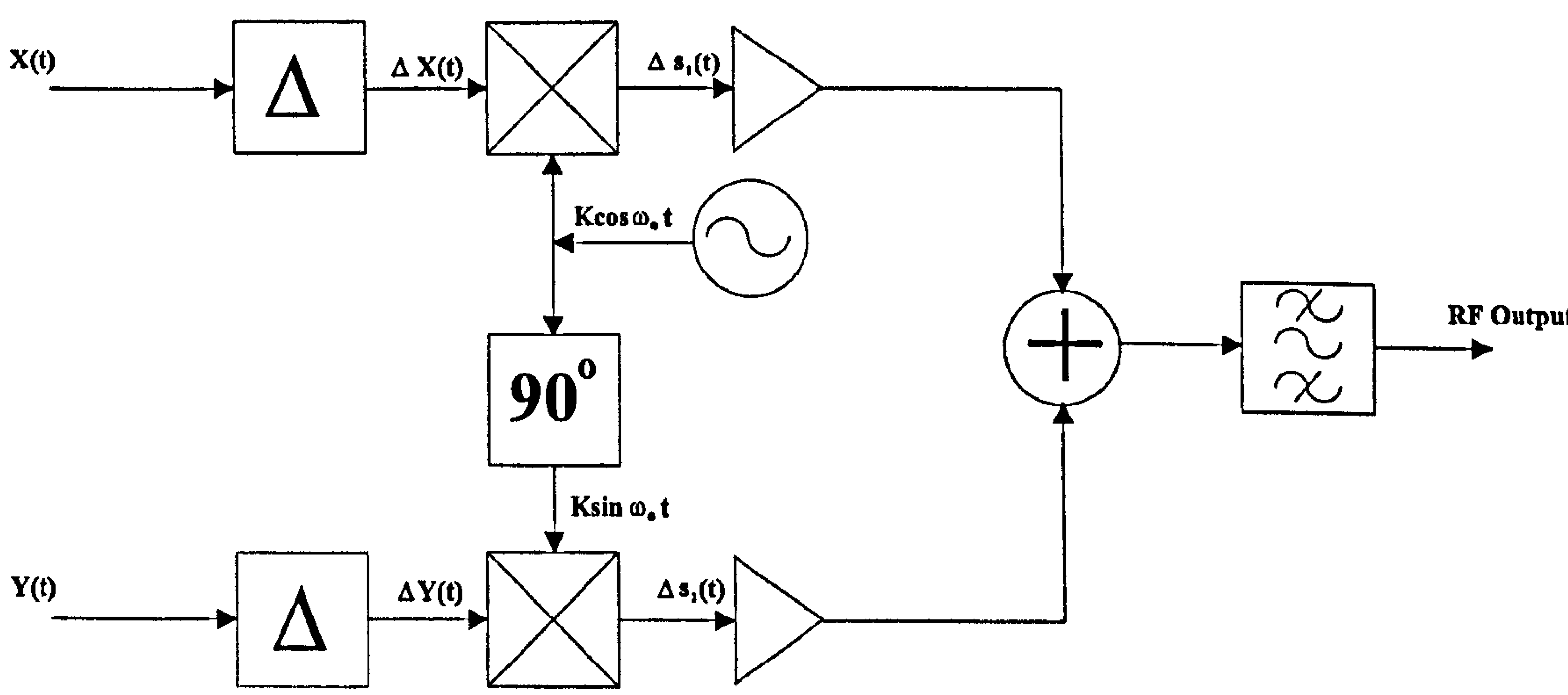

Figure 2.14 The LIST Transmitter

2.5.2 Broadband Techniques

2.5.2.1 Feedforward Linearisation

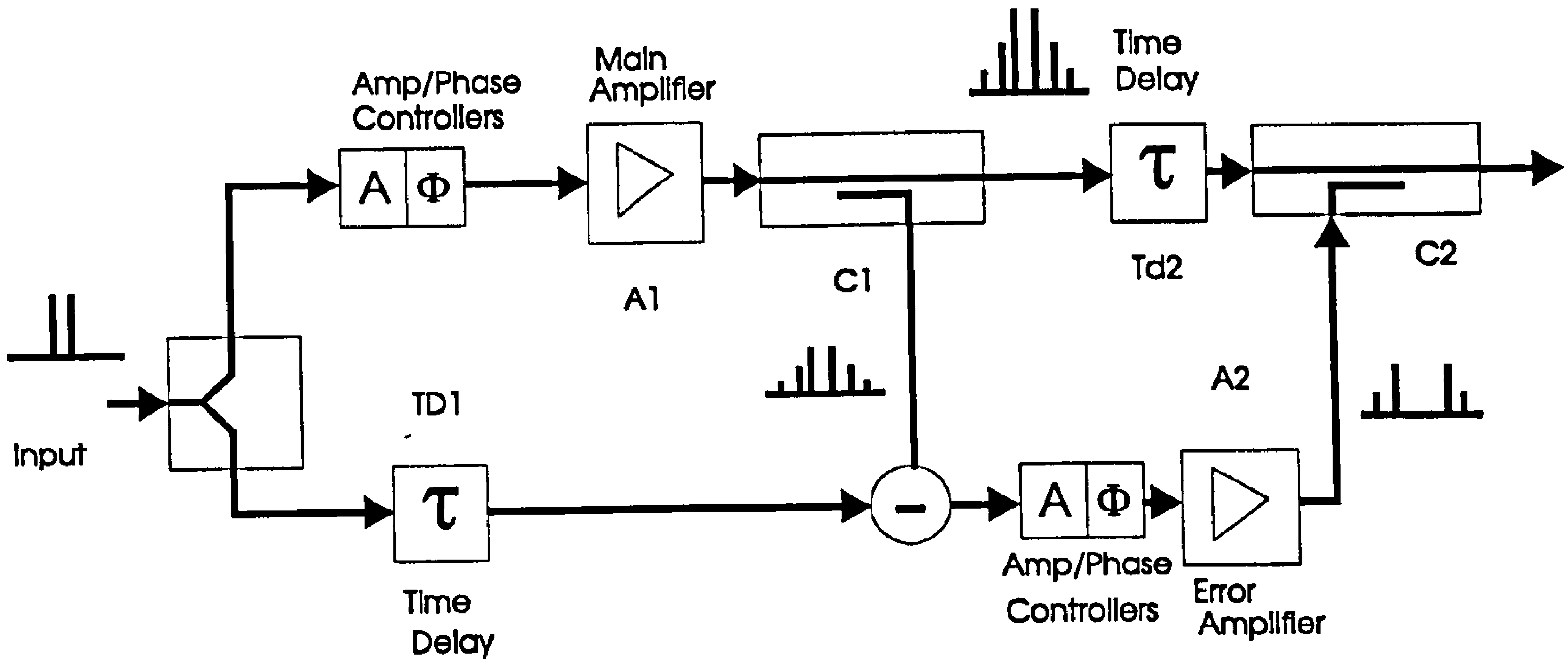

Figure 2.15 The Feedforward Amplifier

The feedforward linearisation technique was first proposed by Black in 1927 [23]. The feedforward technique is shown in figure 2.15. The technique operates in the following manner. Assuming a two tone input the input signal is split into two paths, the main high power path (top) and the lower power error path (bottom). The main path signal is amplified by the main amplifier (A1), this amplifier may be highly non linear and so will introduce distortion. A portion of this distorted signal is tapped via a directional coupler and subtracted for a delayed version of the input signal. This produces an error signal that is mainly made up of distortion products. This error signal is amplified by an error amplifier (A2). This signal is then recombined in anti-phase via a directional coupler with a delayed version of the main amplifier signal. This has the effect of cancelling the main amplifier distortion products at the output of the system. The gain and phase controllers are required to equalise the signals at the two directional couplers C1 and C2 so that maximum cancellation of the input signal and distortion products is achieved. Further distortion cancellation may be achieved with additional feedforward loops, by considering the feedforward amplifier as the main amplifier in another feedforward system.

The technique is open loop and thus unconditionally stable. The technique does however require adaptation of the amplitude and phase in both paths to achieve optimal performance. The linearity performance that can be achieved is excellent and can be maintained over a wide bandwidth. Kenington et al [24] have achieved a reduction in inter-modulation distortion in excess of 40dB over a 30MHz band centred at 900MHz (using two feedforward loops).

2.5.2.2 Postdistortion Linearisation

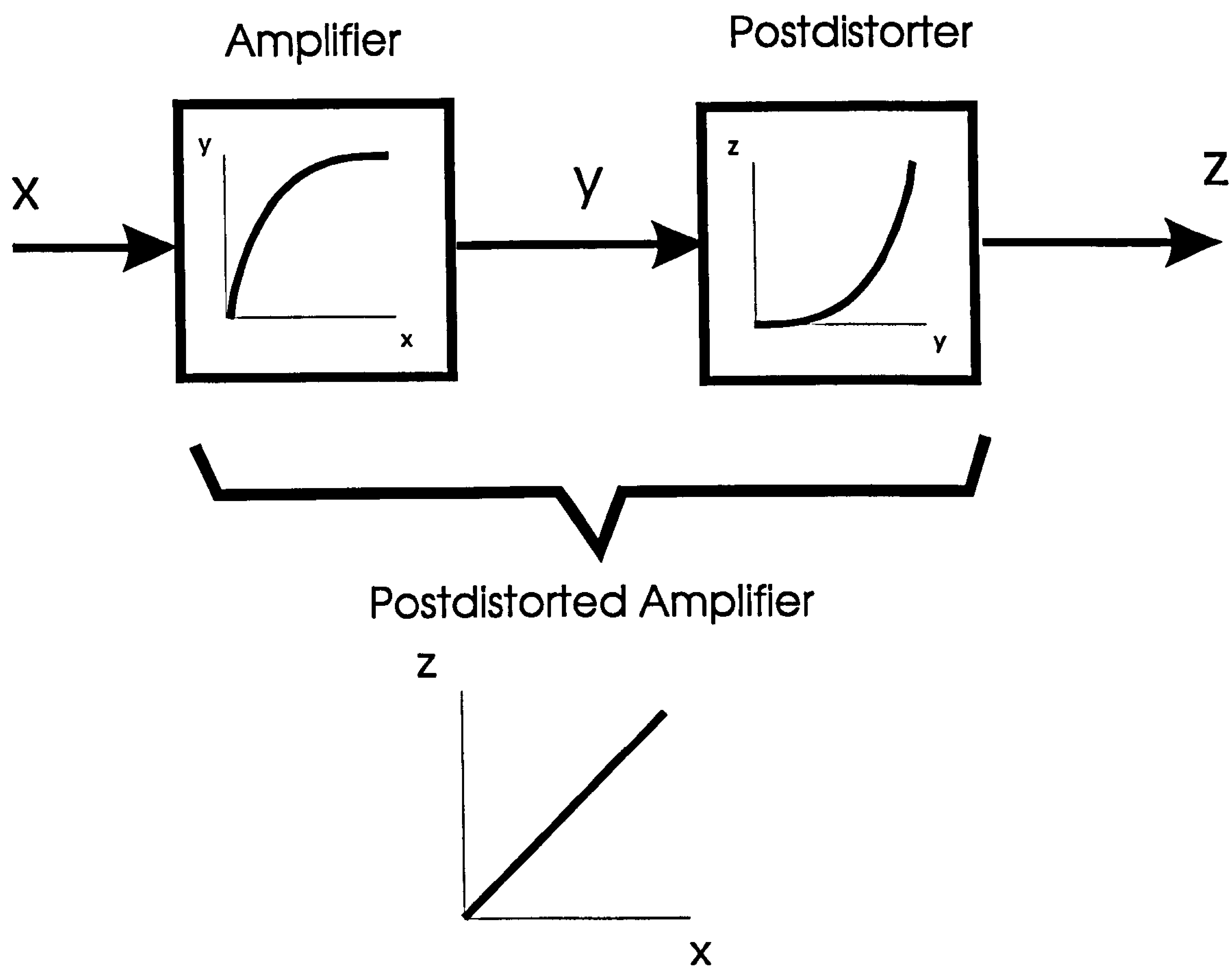

Figure 2.16 Block Diagram of a Postdistorted Amplifier

The postdistortion linearisation technique [25, 26] is similar to predistortion, but the distortion characteristic is implemented after the amplifier rather than before it. The general block diagram form is shown in figure 2.16. The technique is generally unsatisfactory due to the high power handling capability required of the postdistorter. This means that any losses associated with the postdistorter will cause significant degradation of efficiency.

2.5.2.3 Predistortion

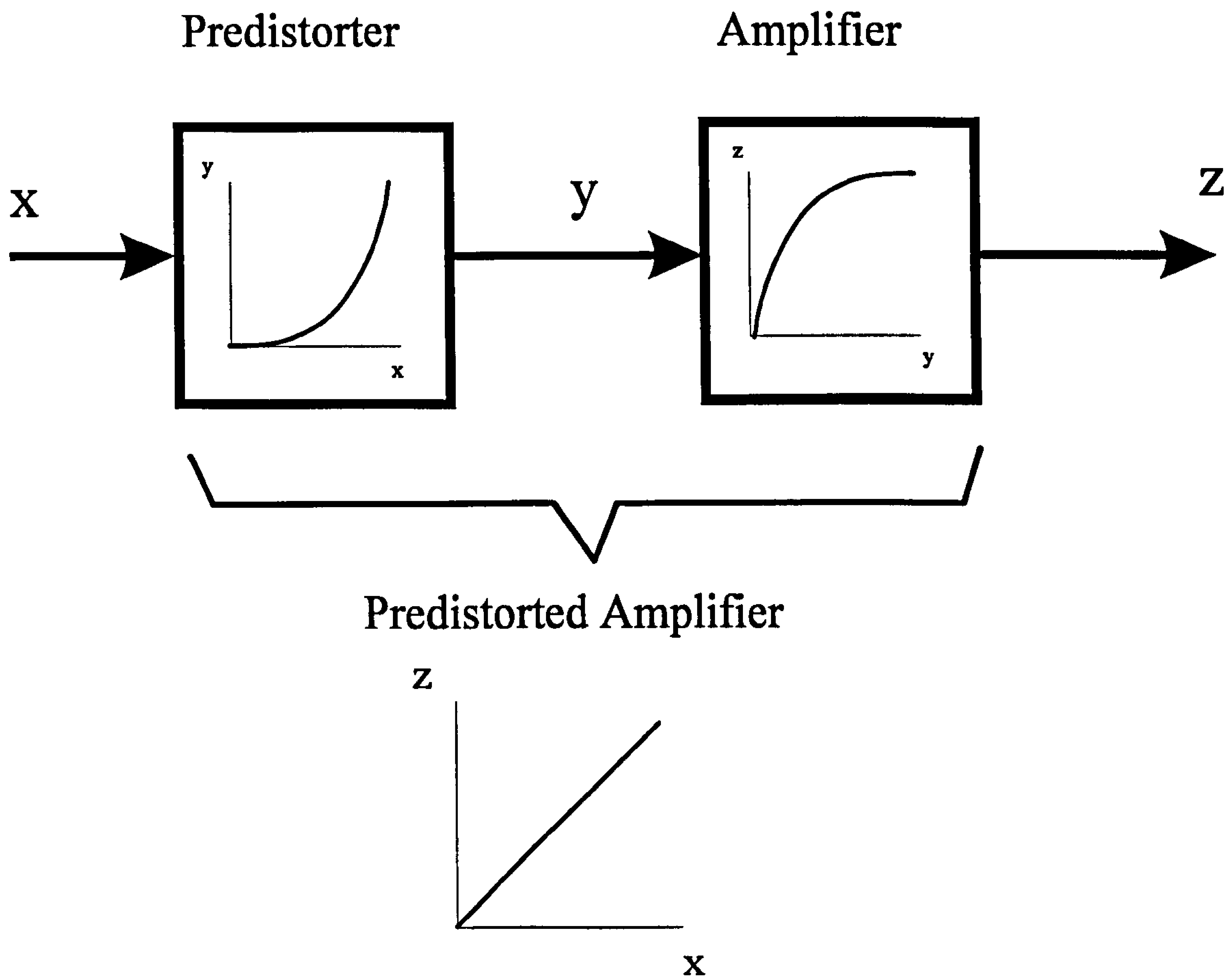

Figure 2.17 Predistorter Block Diagram

Predistortion is a broadband technique which has seen some attention in recent years [27, 28, 29]. This section covers the different types of predistortion and the cancellation that has been achieved in published systems.

Predistortion in its basic form which is shown in figure 2.17 uses a non-linearity at the input to the amplifier that has the inverse of the transfer function of the amplifier being predistorted. The method is essentially open loop and therefore unconditionally stable. Being open loop means that the technique in its simplest form is non-adaptive and so will not respond to changes of amplifier transfer function caused by temperature changes ageing of components or channel switching. Predistortion can be divided into two broad categories, adaptive baseband predistortion and R.F. / I.F. predistortion. Adaptive baseband predistortion has been implemented in [30, 31] using DSP technology, which may be designed to be adaptive. Where as the high frequency techniques represented by R. F. and I. F. predistortion have been implemented using analogue components. These analogue systems are more difficult to design with adaptation built in.

2.5.2.4 R. F. and I. F. Predistortion

R.F. based predistorters work at the desired carrier frequency, whereas I.F. based predistorters work at a reduced frequency and then up-convert the predistorted signal prior to amplification. Both methods use analogue components to produce the predistorted signal. Currently published techniques have employed diodes, amplifiers and FET's as the predistorting elements. These analogue-based predistorters can be broken down into three categories, generic predistorters, piece wise linear predistorters and polynomial predistorters.

2.5.2.5 Generic Predistorters

Generic based predistorters [32, 33] use a non-linear device which has a transfer characteristic which is the inverse of the amplifier which is being linearised. Generic predistortion has thus far been used on amplifiers with a compressive characteristic such as class AB television amplifiers and TWT satellite amplifiers. So the devices used to date have exhibited expansive characteristics to counteract the compressive nature of the amplifiers. Therefore devices used as predistorters so far include dual gate GaAs FETs operating close to pinch-off and Schottky diodes.

2.5.2.6 Piece Wise Linear Predistortion

The piecewise linear predistorter is a variation on the generic predistorter which for increased complexity increases the number of degrees of freedom, providing improved flexibility and linearity. The method uses a non-linear network which approximates the ideal predistorter characteristic using a number of segments. The method is completely general and may be used to linearise any kind of non-linearity. Piecewise linear predistortion (PLP) approximates the ideal predistorter characteristic using a number of piecewise linear segments [34, 35, 36], this arrangement is shown in figure 2.18.

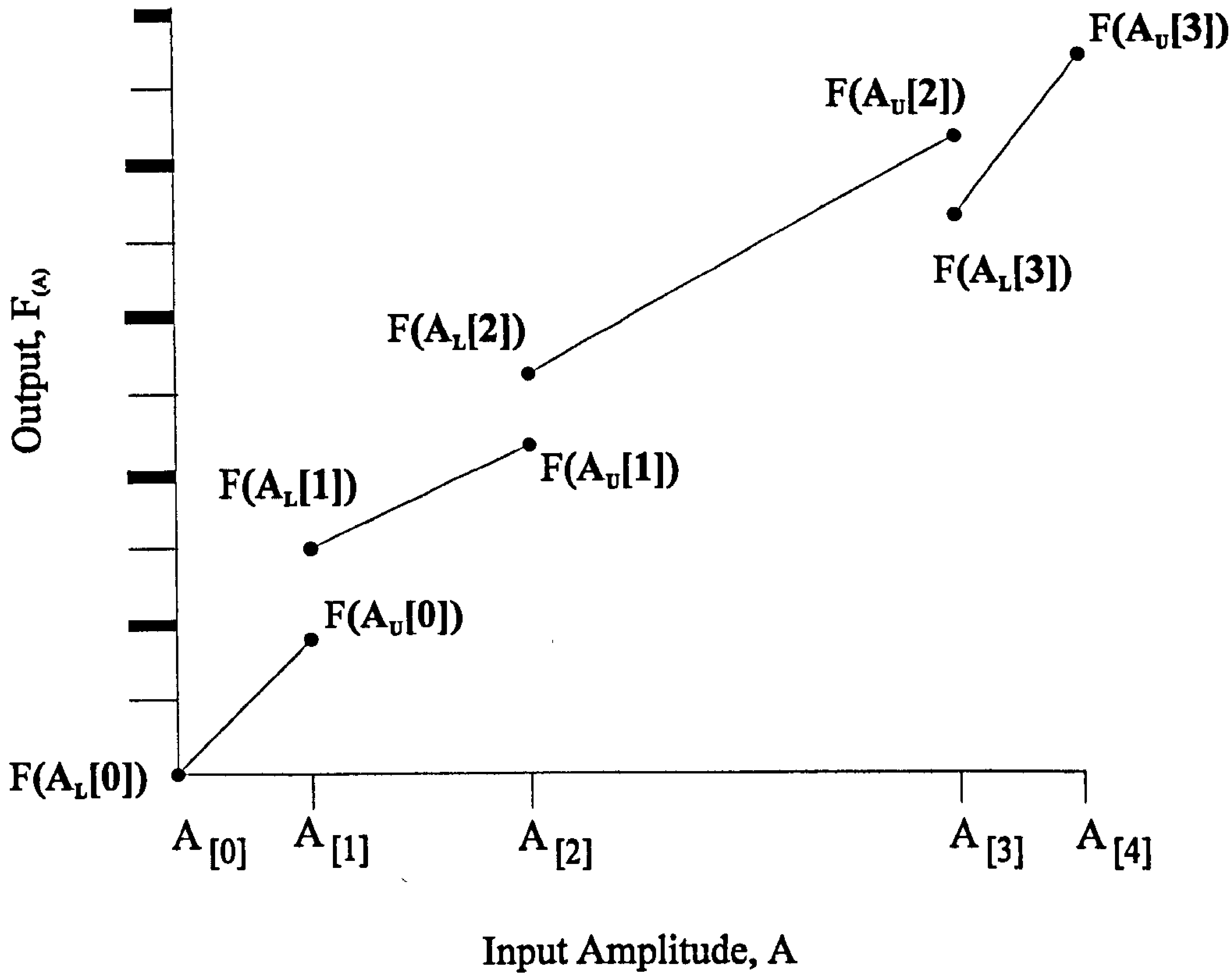

Figure 2.18 Discontinuous Piecewise Linear Predistorter

The transition between the segments occurs at what is known as a knot. The characteristic of a general PLP is a discontinuous function. It is also possible to constrain the characteristic to be continuous forming the continuous first order PLP. This method has the advantage of using relatively simple circuit architectures to implement the required functions.

2.5.2.7 Polynomial Predistortion

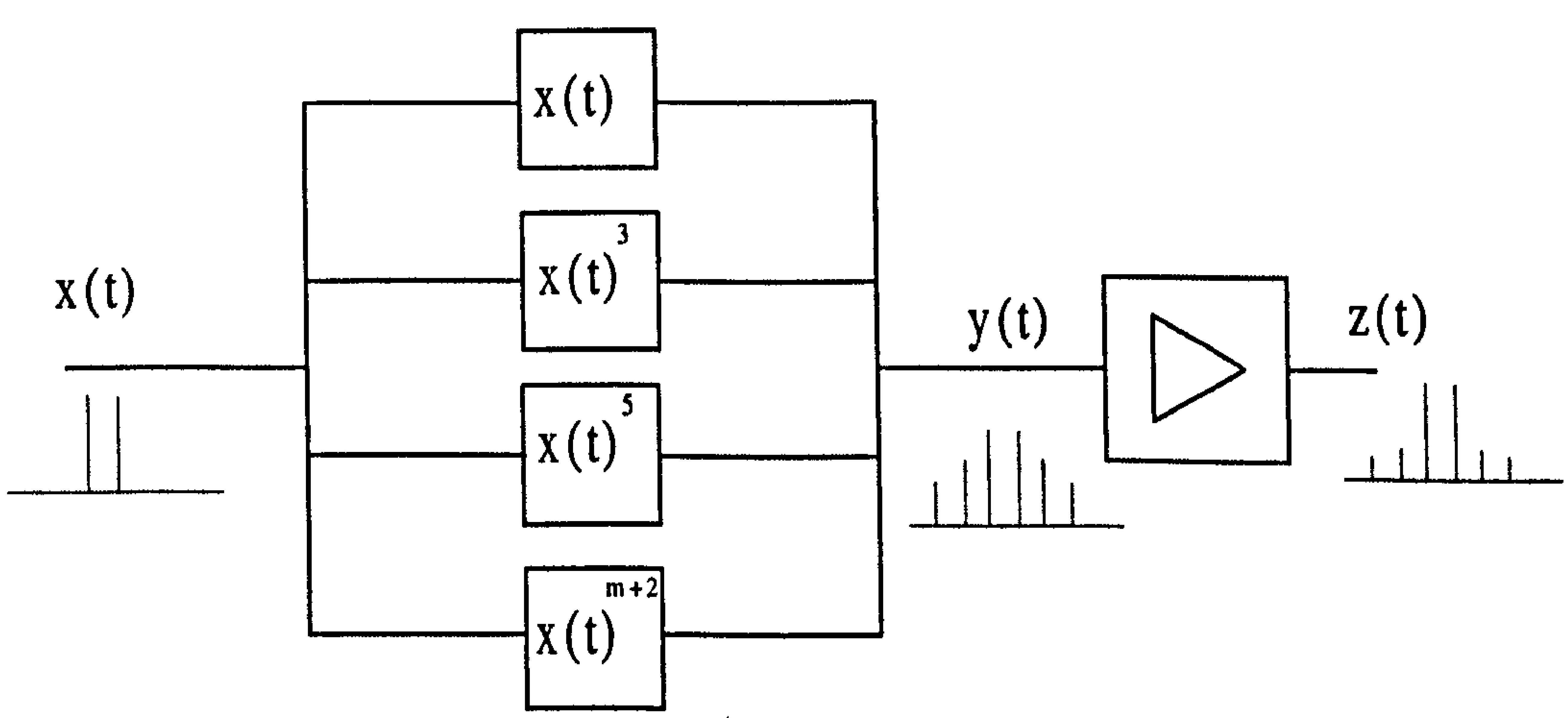

Figure 2.19 Polynomial Predistorter

Polynomial predistortion [37, 38, 39, 40] shown in figure 2.19 is a technique used to cancel a single order of non-linearity. The predistorters operate by applying a single order of non-linearity to the amplifier with the effect of reducing the level of the IMD product, which is the same order as the predistorter. So a cubic non-linearity can be used to reduce the third-order IMP's and a quintic non-linearity can be used to reduce the fifth-order IMP's. The predistorter operates by vector addition of the intermodulation products from the predistorter with the intermodulation products generated by the amplifier. The predistorter produces intermodulation products which are of equal amplitude and in anti-phase with the products generated by the amplifier, the effects of this cancellation may be seen in figure 2.19. This method can be applied to any further order of non-linearity such as 7th, 9th etc. The polynomial predistorter although it will only correct one order of non-linearity, has the advantage of providing a cheap and simple form of linearisation for amplifiers with single order non-linearities, i.e. amplifiers that have a mainly 3rd cubic transfer function. TWT amplifiers, which are frequently used in satellite systems, have been linearised with polynomial predistorters. Polynomial predistorters are also used to linearise class AB amplifiers for television systems. Chapters 4 and 5 investigate polynomial predistorters in more detail and develop a new type of polynomial predistorter.

2.5.3 Hybrid Linearisation Schemes

The use of hybrid linearisation schemes has received limited attention in the literature. These schemes attempt to overcome the efficiency or bandwidth limitations of more traditional methods of linearisation. These methods have shown promise to date due to their potential for greatly enhancing amplifier performance. To date two methods have been proposed, they are the use of predistortion combined with cartesian loop (Composite Modulation Feedback Transmitter) and predistortion combined with feedforward.

2.5.3.1 Composite Modulation Feedback Transmitter

Composite modulation feedback was proposed by Mansell [41] with the intention of overcoming some of the limitations of the cartesian loop technique. A block diagram of this technique is shown in figure 2.20. The technique operates in the following way; the input signal is first predistorted using a purely cubic non-linearity, this predistorted input signal is then correctly scaled and then coherently summed with the Cartesian loop error signal. The effect of this process is to effectively make the Cartesian loop operate around a partially corrected non-linearity, this has several advantages for the predistorter and Cartesian loop design. The simple predistortion design used can add an additional 10 to 20dB of linearity improvement using a very low resolution (4 – 8 bit) data conversion; the Cartesian loop corrects any low level distortion remaining after the predistortion process. The advantage for the Cartesian loop process is the reduction in loop gain required for an equivalent combined linearity improvement. This has the added advantage of improving loop stability and flexibility that significantly reduces transmitter wideband noise. The system is set up such that the Cartesian loop is running with a low loop gain, resulting in an IMP improvement of 25dB. The predistorter reduces the IMP's by a further 11dB, which results in all IMP's being below -67dBc. These results were obtained with a tone spacing of 20kHz. The main applications of this technique are narrowband basestation transmitter design and mobile transmitters using power efficient, nonlinear R.F. amplifiers.

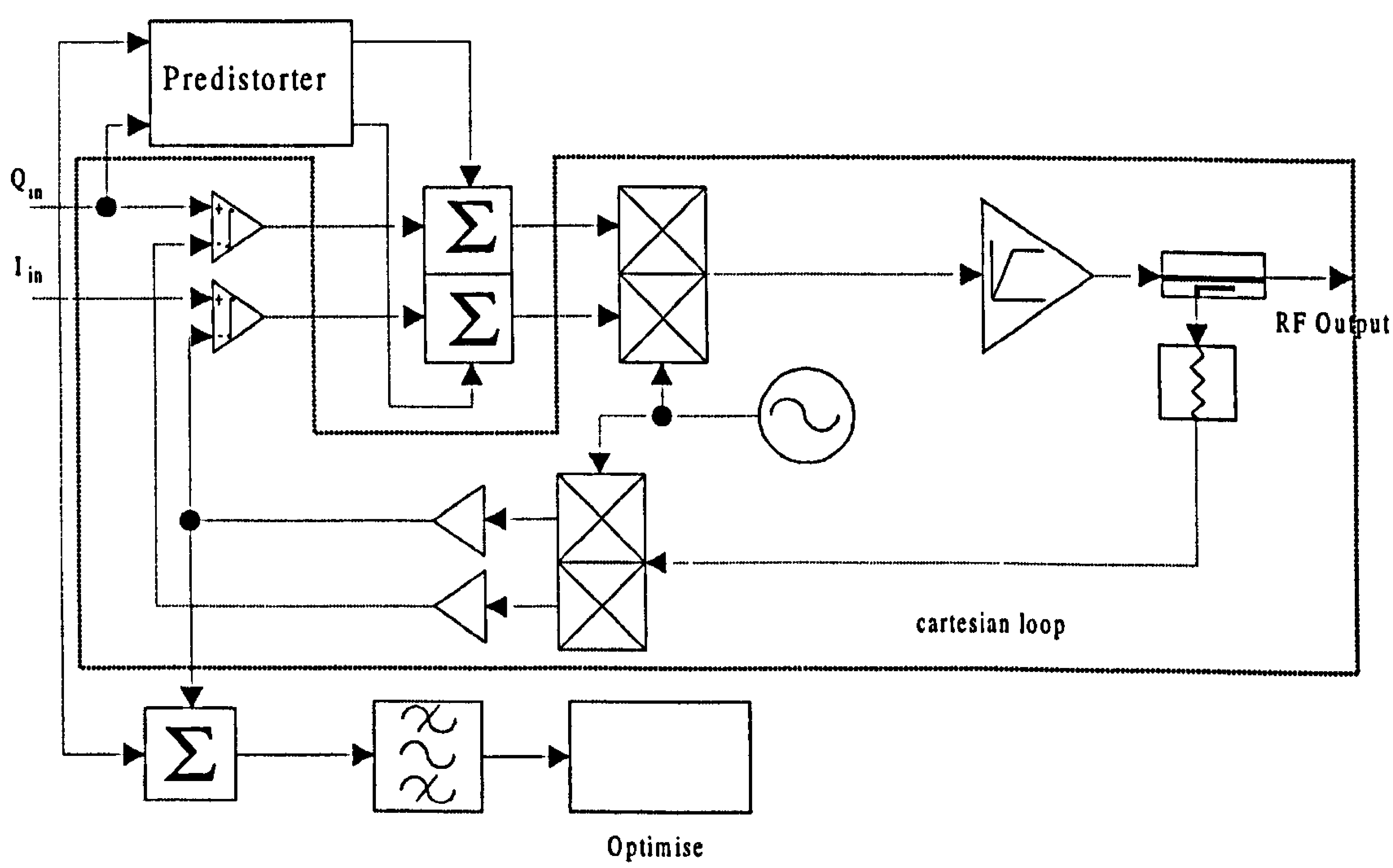

Figure 2.20 Composite Modulation Feedback Transmitter

2.5.3.2 Predistortion and Feedforward

The combination of feedforward and predistortion techniques was first proposed in [42]. The aim of this method is to improve the efficiency of feedforward amplifiers without a loss in linearity performance. The method operates in the following manner, a predistortion network is fitted before the main and if desired the error amplifier this arrangement is shown in figure 2.21. This predistorter provides a limited amount of linearity improvement over the raw amplifier performance. The feedforward process then obtains the remaining linearity improvement. The overall effect of this system is an increase in linearisation performance with a system efficiency that is greater than the feedforward process alone. Results obtained in [43] show that 29dB of IMP improvement is possible for an amplifier with an overall efficiency of 21%. These results were obtained using class C amplifiers for the main and error amplifiers.

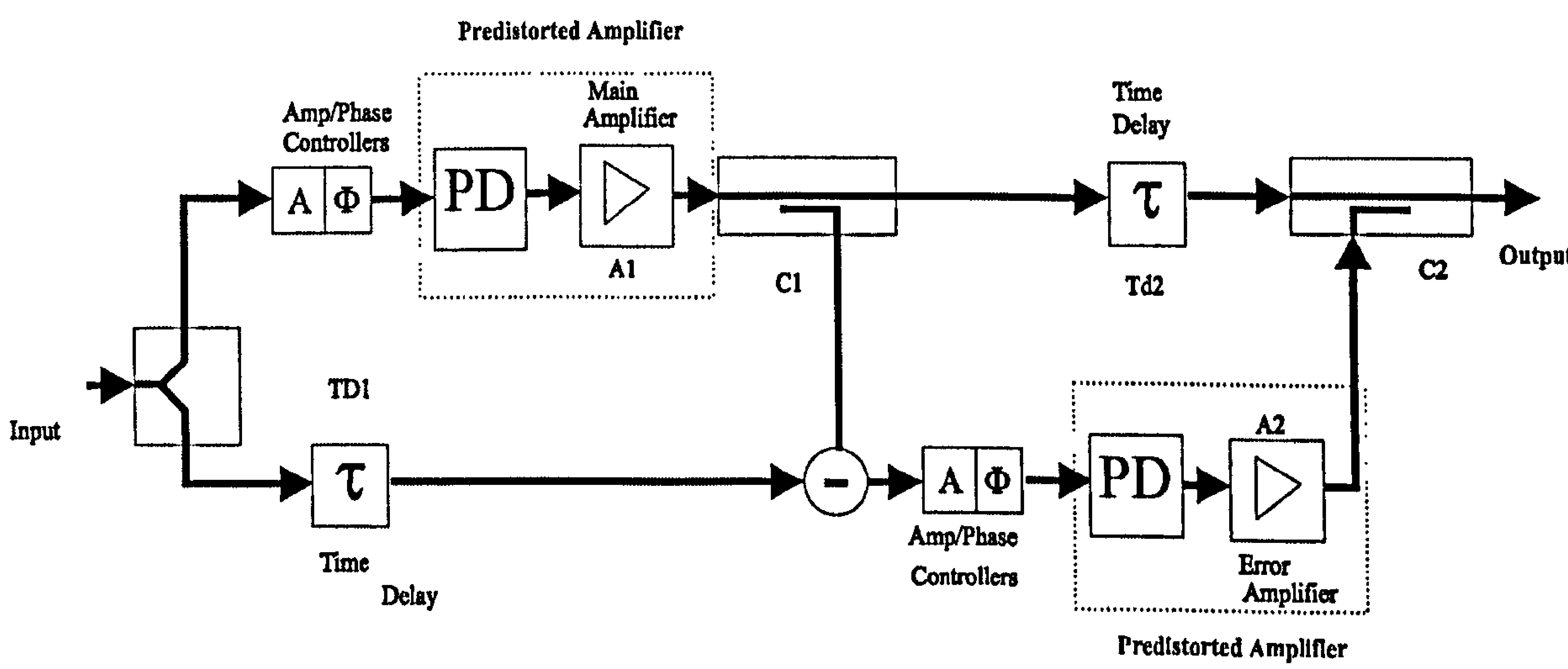

Figure 2.21 Predistortion and Feedforward Amplifier

It has been shown by Parsons [44] that the efficiency of a practical feedforward amplifier may be expressed as:

$$\eta_{ff} = \frac{\eta_{A1}\eta_{A2}C_{DC}(1 - C_{DC})}{\eta_{A1}F_{IM}(1 - C_{DC}) + \eta_{A2}LC_{DC}(1 + F_{IM})} \quad (2.15)$$

Where: η_{A1} is the main amplifier efficiency

η_{A2} is the error amplifier efficiency

C_{DC} is the output coupler coupling factor

F_{IM} is the fractional power of a single IMP

L is the fractional loss of signals through the delay

It has also been shown in [44] that there is an optimum value of coupling factor $C_{DC,OPT}$ which is given by:

$$C_{DC,OPT} = \frac{\eta_{A1} F_{IM} \pm \sqrt{\eta_{A1}\eta_{A2} F_{IM} L(1+F_{IM})}}{\eta_{A1} F_{IM} - \eta_{A2} L(1+F_{IM})} \qquad (2.16)$$

These equations may be used to analyze the effect that changes in amplifier efficiency and the IMP performance that the amplifier that is being linearised have on the efficiency of the feedforward amplifier being linearised. For the purposes of illustrating the benefit that a predistorter has on the efficiency of a feedforward amplifier two cases have been considered; main and error amplifier efficiencies of 15% which is shown in figure 2.22 and main and error amplifier efficiencies of 60% which is shown in figure 2.23. In both cases an optimum coupling factor has been assumed. Figure 2.22 shows that the efficiency of the feedforward amplifier is dominated by the delay loss and not the reduction in IMP introduced by the predistortion process. However the situation in figure 2.23 is somewhat different when class C amplifiers are used as the main and error amplifiers. It maybe seen that significant benefits maybe obtained with the use of predistortion, for example with 10dB of cancellation introduced by the predistortion system the efficiency has been increased by 10% assuming a 1dB delay loss. This technique therefore potentially has the capability of making broadband efficient linear amplification a reality.

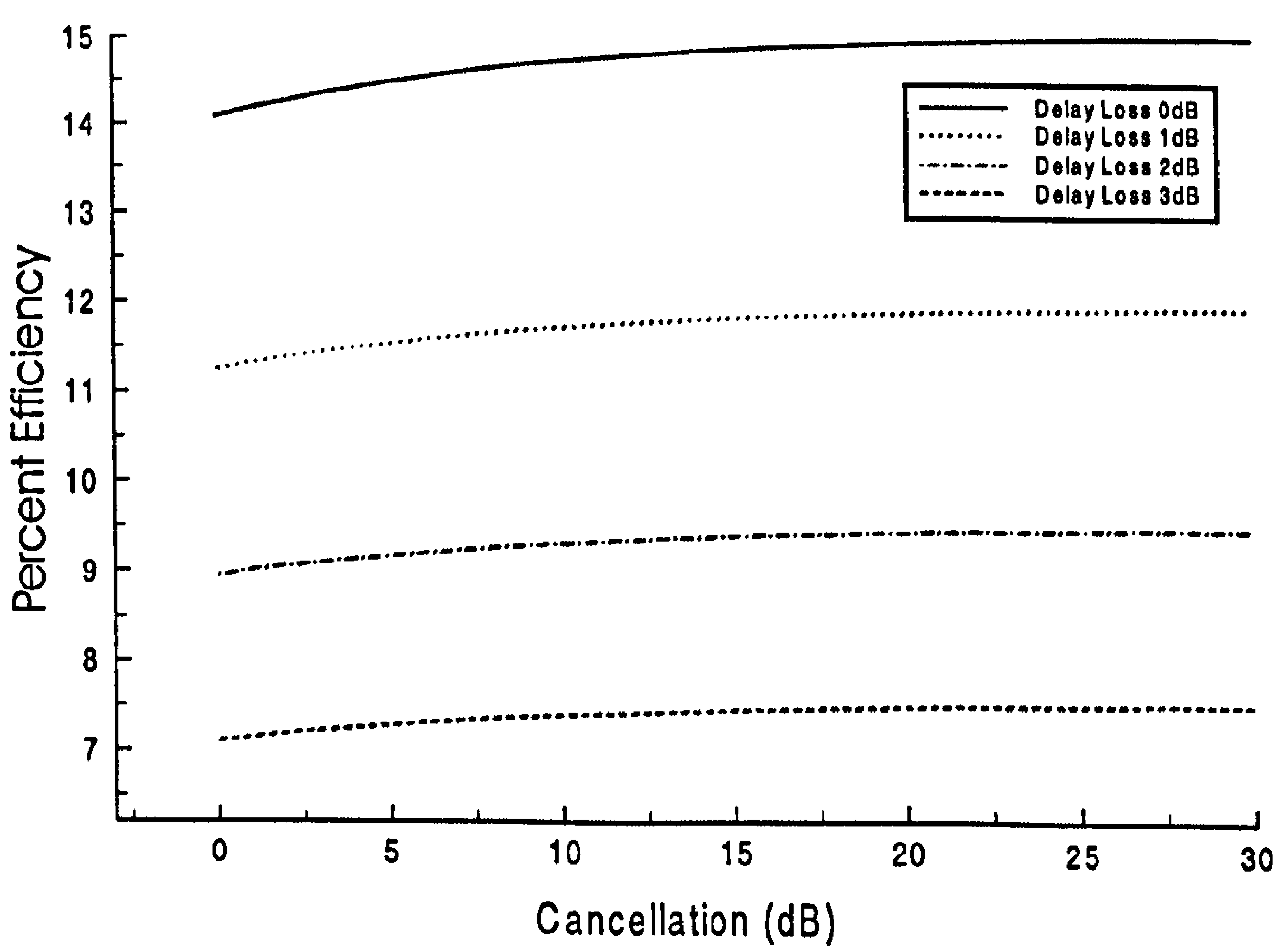

Figure 2.22 Combined Amplifier Efficiency with 15% Efficient Main and Error Amplifiers

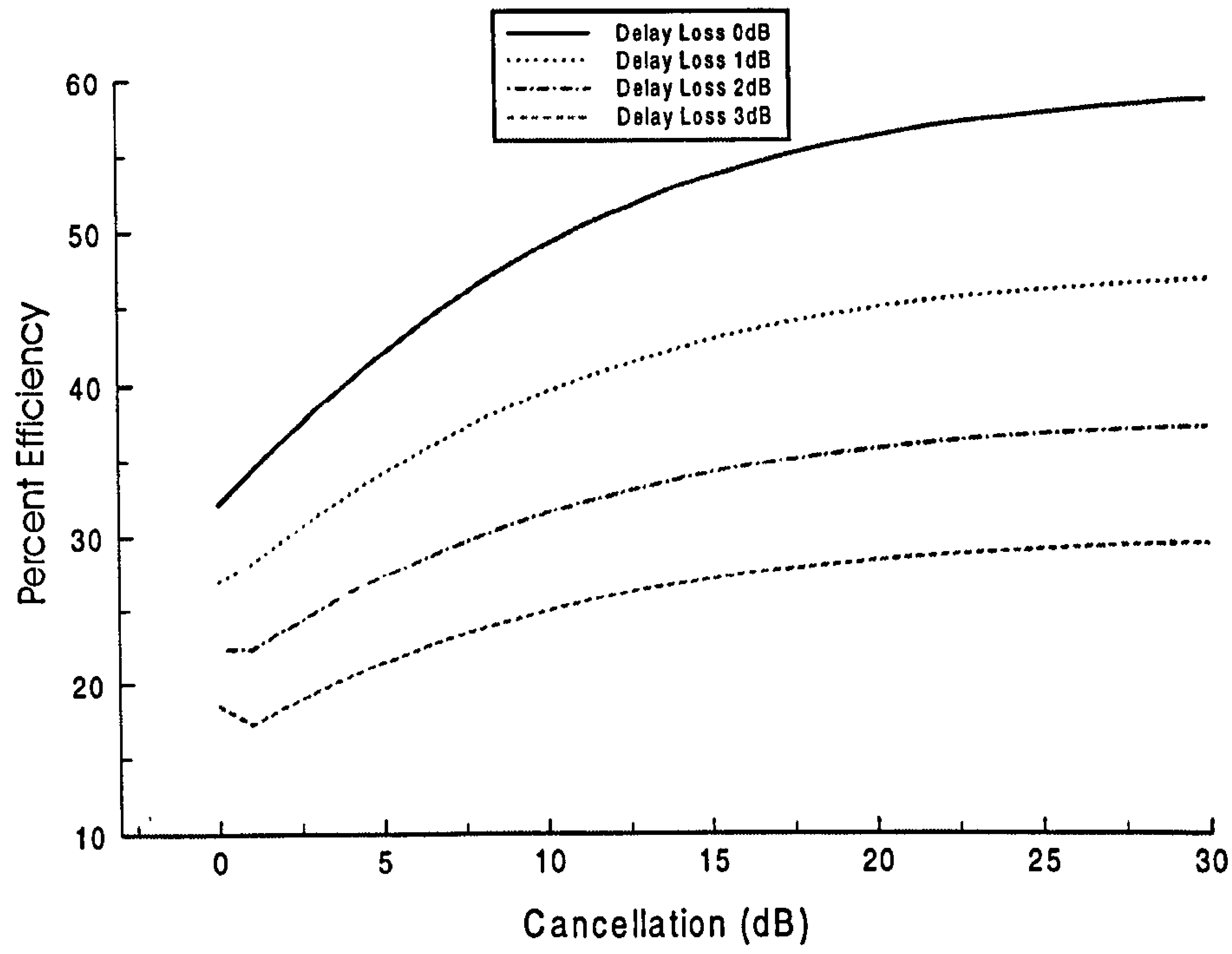

Figure 2.23 Combined Amplifier Efficiency with 60% Efficient Main and Error Amplifiers

2.6 Discussion

This chapter has explained the main types of non-linearity encountered in amplifiers. The effects of amplitude distortion, intermodulation distortion and AM to PM conversion. The theoretical background to these effects has been covered and the relationship this has to practical examples. In an ideal situation amplifiers would have no AM to AM or AM to PM distortion which would mean no intermodulation distortion. In practice however all amplifiers have some intermodulation distortion. The purpose of this research is to find new ways of reducing this distortion to a minimum while maintaining amplifier efficiency. This chapter has also introduced same methods for improving amplifier linearity, which have been tried in the past. Some of these linearisation techniques are now being applied to commercial systems.

This chapter has also shown that feedforward and predistortion are currently the only two methods available for the linearisation of broadband power amplifiers. Although feedforward has limitations on the efficiency that maybe achieved, it has been shown that by combining feedforward with predistortion techniques then it is theoretically possible to have efficient linear power amplification. The remainder of the thesis is aiming at developing methods for broadband linearisation of power amplifiers using predistortion techniques.

References

[1] Tri. T. Ha, Solid State Microwave Amplifier Design, John Wiley & Sons 1981.

[2] Herbert. L. Kraus, Charles. W. Bostain and Frederick. H. Raab, Solid State Radio Engineering, John Wiley & Sons New York, 1980.

[3] F.E. Terman, Electronic and Radio Engineering, McGraw-Hill Book Company, New York 1955.

[4] P.J. Baxandall, Transistor Sinewave Oscillators, Proc. IEE (London), 106, Part B, Supplement. 16, pages 748 – 758, May 1959.

[5] N.O. Sokal and A.D. Sokal, Class E, a New Class of High-Efficiency Tuned Single-Ended Power Amplifiers, IEEE Journal of Solid State Cicuits, SC-10, pages 168 – 176, June 1975.

[6] B.D. Bedford, Electric Amplifying Circuits, U.S. Patent 1,874,159, 30th August 1932.

[7] A. Mitchell, A 135 MHz feedback amplifier, IEE Colloquium on Broadband High Frequency Amplifiers - Practice and Theory (London, U. K.), pages 2.1 - 2.6, November 1976.

[8] R. Wilkinson, J. MacLeod, M. Beach and A. Bateman, Linear transmitter design for MSAT terminals, IEEE 41st Vehicular Technology Conference, volume 1, pages 297 – 301, June 1990.

[9] A. Bateman and D. Haines, Direct conversion transceiver design for compact low-cost portable mobile radio terminals, IEEE 39th Vehicular Technology Conference , pages 1/57 – 1/62, May 1989.

[10] M. Johansson and T. Mattsson, Linearised high-efficiency power amplifier for PCN, Electronics Letters, volume 27, pages 762 – 764, April 1991.

[11] V. Petrovic and W. Gosling, Polar-Loop Transmitter, Electronics Letters, volume 15 pages 286 – 287, May 1979.

[12] Yoshinari Nagata, Linear Amplification Technique for Digital Mobile Communications, IEEE Vehicular Technology Conference, volume 1, pages 159 – 164, May 1989.

[13] Andrew S. Wright and Willem G. Durtler, Experimental Performance of an Adaptive Digital Linearized Power Amplifier, IEEE Transactions on Vehicular Technology, volume 41, No. 4, pages 395 – 400, November 1992.

[14] A. Baranyi, T. Mesterhazy and L. Zakarias, Baseband Equalization - A Novel Method for Intermodulation Noise Reduction in Microwave FM Systems, IEEE National Telecommunications Conference, pages 3.2.1 – 3.2.4, 1981.

[15] L. R. Kahn, Single-Sideband Transmission By Envelope Elimination and Restoration, Proceedings of the IRE, volume 40, pages 803 – 806, July 1952.

[16] L. R. Kahn, Comparison of Linear Single-Sideband Transmitters with Envelope Elimination and Restoration Single-Sideband Transmitters, Proceedings of the IRE, volume 44, pages 1706 – 1712, December 1956.

[17] D. C. Cox, Linear Amplification with Nonlinear Components, IEEE Transactions on Communications, volume 22, pages 1942 – 1945, December 1974.

[18] S. A. Hetzel, Power Efficient Linear Transmitters and the LINC Technique, University of Bristol, February 1993.

[19] A. Bateman, The Combined Analogue Locked Loop Universal Modulator (CALLUM), IEEE Vehicular Technology Conference, volume 1, pages 759 – 763, May 1992.

[20] K. Chan and A. Bateman, Linear Modulators Based on RF Synthesis: Realization and Analysis, IEEE Transactions on Circuits and Systems -I:Fundemential Theory and Applications, volume 42, pages 321 – 333, June 1995.

[21] D. J. Jennings and J. P. McGeehan, A high-efficiency RF transmitter using VCO-derived synthesis: CALLUM, IEEE Radio and Wireless Conference, pages 137 – 140, August 1998.

[22] D. C. Cox, Linear Amplification by Sampling Techniques: A New Application for Delta Codes, IEEE Transactions on Communications, volume 23, pages 793 – 798, August 1975.

[23] H. S. Black, U. S. Patent 1,686,792, October 1929.

[24] P. B. Kenington, R. J. Wilkinson and J. D. Marvill, A Multi - Carrier Amplifier for Future Mobile Communications Systems, IEE International Conference on Mobile Radio and Personal Communications, No. 6, pages 151 – 162, December 1991.

[25] A. Prochazka and P. Lancaster and R. Neumann, Amplifier Linearization by Complimentary Pre or Post-Distortion, IEEE Transactions on Cable Television, volume CATV-1, No. 1, pages 31 – 39, October 1976.

[26] S. P. Stapleton and Le Quach, Reduction of Adjacent Channel Interference using Postdistortion, IEEE Vehicular Technology Conference, volume 2, pages 915 – 918, May 1992.

[27] Shigenobu Aihara, Toshio Nishiumi, Yoshinori Fujiki and Sachiro Fukuda, GaAs FET Power Amplifiers as Substitutes for TWT Amplifiers in A Multilevel QAM Digital Radio System, ICC, volume 1, pages 6 – 10, June 1987.

[28] Hideo Ashida, Yoshiaki Suzuki, Isamu Umino and Norio Tozawa, C Band 100 Watts GaAs FET Amplifier for Digital Radio, ICC, volume 1, pages 23 – 27, June 1987.

[29] H. Balshem and B. Hewitt and H. J. Thaler, Advanced RF Power Amplifiers for High Capacity Digital Radio Systems, ICC, volume 1, pages 28 – 33, June 1987.

[30] James K. Cavers, Amplifier Linearization Using a Digital Predistorter with Fast Adaptation and Low Memory Requirements, IEEE Transactions on Vehicular Technology, volume 39, No. 4, pages 374 – 382, November 1990.

[31] Michael Faulkner and Matts Johansson, Adaptive Linearization Using Predistortion - Experimental Results, IEEE Transactions on Vehicular Technology, volume 43, No. 2, pages 323 – 332, May 1994.

[32] D. Cahana, J. R. Potukuchi, R. G. Marshalek and D. K. Paul, Linearized Transponder Technology for Satellite Communications Part 1 Linearizer Circuit Development and Experimential Characterization, Comsat Technical Review, volume 15, No. 2A, pages 277 – 306, August 1985.

[33] G. Eggers and R. Kohl, 2GHz Bandwidth Predistortion Linearizer for Microwave Power Amplifiers at Ku-Band, 24th European Microwave Conference, volume 2, pages 1501 – 1505, September 1994.

[34] Ph Charas and J. D. Rogers, Improvements in On - Board TWTA Performance using Amplitude and Phase Predistortion, IEE International Conference on Satellite Communication System Technology ICP 126, pages 270 – 280, April 1975.

[35] M. T. Abuelma'atti, Reduction of intermodulation in communications amplifiers by pre-correction techniques, University of Bradford, 1979.

[36] K. P. Parsons, Highly Efficient Broadband Linearisation of R. F. Amplifiers for Satellite and Cellular Applications, University of Bristol, March 1996.

[37] M. Gahderi, S. Kumar and D. E. Dodds, Adaptive predistortion lineariser using polynomial functions, IEE Proceedings on Communications, volume 141, No. 2, pages 49 – 55, April 1994.

[38] P. Bura, D. Gelerman and P. Ntake, Linear Solid State Power Amplifier for 64 QAM Radios, ICC, volume 1, pages 1 – 5, June 1987.

[39] Nobuaki Imai, Toshio Nojima and Takehiro Murase, High Power Linear Amplifier Linearization for a Multi-Carrier Digital Microwave System, IEEE Globecom, pages 547 – 553, 1986.

[40] Surinder Kumar, Power Amplifier Linearization using MMICs, Microwave Journal, pages 96 – 104, April 1992.

[41] A. R. Mansell and A. Bateman, Transmitter linearisation using composite modulation feedback, Electronics Letters, Vol 32, pages 2120 – 2121, November 1996.

[42] P. B. Kenington, K. J. Parsons and D. W. Bennett, Broadband linearisation of high-efficiency power amplifiers, Proceedings of the Third International Mobile Satellite Conference, pages 59 – 64, June 1993.

[43] K.J. Parsons, R.J. Wilkinson and P.B. Kenington, A Highly Efficient Linear Amplifier for Satellite and Cellular Applications, Globecom , pages 203 – 207, November 1995.

[44] K. J. Parsons and P. B. Kenington, The Efficiency of a Feedforward Amplifier with Delay Loss, IEEE Transactions on Vehicular Technology, Vol 43, No. 2, pages 407 – 412, May 1994.

Chapter 3
Polynomial Predistortion

The following chapter deals with the mathematical theory of predistortion and the mathematical analysis of an ideal polynomial predistorter.

3 Polynomial Predistortion

3.1 Introduction

The review of the current linearisation schemes carried out in chapter 2 has shown that predistortion along with feedforward are currently the only techniques with the capability to linearise a broadband RF power amplifier. Research has been on going for a number of years investigating the different forms of predistortion technique that may be applied to RF power amplifiers. The different forms of predistortion, which have been implemented, are reviewed in chapter 2. When considering a form of predistortion for the broadband linearisation of amplifiers the problems highlighted in chapter 2 must be born in mind and so having considered these factors only feedforward, RF and IF predistortion have the capability to perform broadband linearisation of RF power amplifiers.

The use of polynomial predistortion has been wide spread for many years, particularly in the fields of satellite power amplification and high bit-rate point-to-point links [1, 2]. This form of amplifier linearisation has traditionally been used where only modest degrees of linearity improvement are required, due to the poor matching usually achieved between the predistorter and amplifier characteristics. More recently, however, the use of predistortion as an aid to the feedforward correction process has been suggested [3] and systems are currently being fabricated with purpose designed silicon for both the PA device and the predistorter. This paves the way for R.F. predistortion to achieve far greater levels of linearity improvement. Because of the open loop nature of this type of predistortion the potential for good broadband performance is excellent. To date to the best of the author's knowledge no analysis has been undertaken to predict the necessary gain and phase matching for an ideal polynomial predistorter.

3.2 The Ideal Predistorter

An ideal predistorter should in theory provide the exact inverse transfer function of the amplifier being linearised. Considering a block diagram of a general predistortion system which is shown in figure 3.1

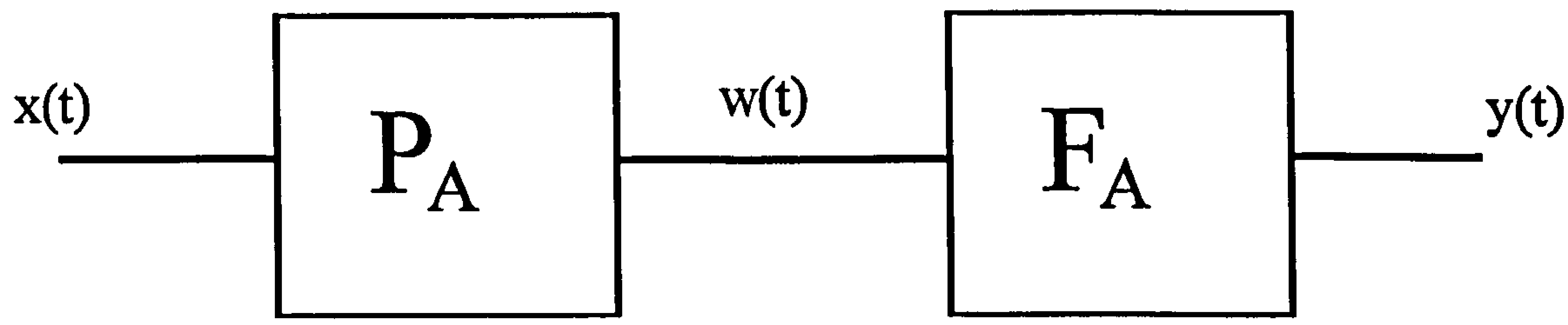

Figure 3.1 A General Predistortion System

The operation of the system maybe expressed as follows:

If an input signal $x(t)$, is applied to the predistorter which has a transfer function of $P_A(\bullet)$, then the resulting output may be expressed as $w(t)$

$$w(t) = P_A(x(t)) \tag{3.1}$$

This signal is then applied to the amplifier $F_A(\bullet)$ being predistorted which generates an output $y(t)$

$$y(t) = F_A(w(t)) \tag{3.2}$$

Which may also be expressed in terms of the transfer functions of the predistorter and the amplifier, the result being

$$y(t) = F_A(P_A(x(t))) \tag{3.3}$$

For the case of an ideal linear amplifier the output will be the input multiplied by a constant k i.e. $y(t) = kx(t)$, therefore the required predistorter characteristic is given by

$$P_A(x(t)) = F_A^{-1}(kx(t)) \tag{3.4}$$

Therefore the ideal predistorter transfer characteristic is the inverse of the amplifier transfer characteristic. This analysis may be extended to cover the AM to AM and AM to PM distortion [4] for an amplifier which generates AM to AM distortion $F_A(\bullet)$ and AM to PM distortion $G_A(\bullet)$ it may be shown that

$$P_A(A) = F_A^{-1}(kA) \tag{3.5}$$

and

$$G_P(A) = \theta - G_A(P_A(A)) \tag{3.6}$$

Equation 3.6 may also be expressed in the following form

$$G_P(A) = \theta - G_A\left(F_A^{-1}(kA)\right) \tag{3.7}$$

So in order to obtain a linear amplifier from a predistorter and non-linear amplifier combination the predistorter has to map exactly the inverse of the AM to AM and the AM to PM characteristics. The above analysis applies to a general predistortion system. For more specific cases the analysis must be carried out for the actual predistorter/amplifier combination being used. The next section carries out an analysis for an ideal cubic polynomial predistorter which is used to linearise an amplifier which is 3rd order non-linear which will be the case for a typical class A amplifier or a class AB amplifier which is biased mainly for class A operation [5].

3.3 Derivation of Optimal Polynomial Predistorter Coefficients

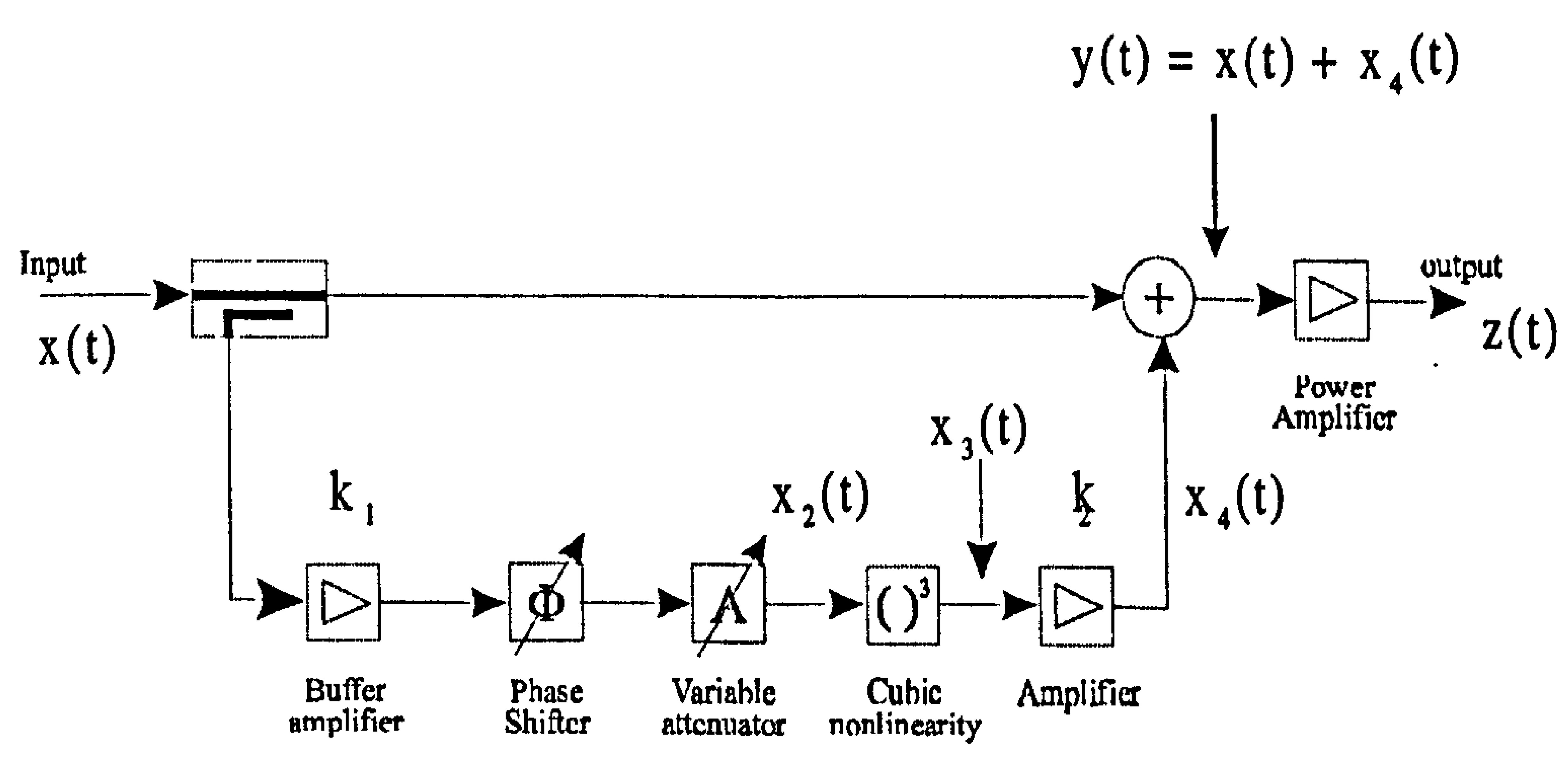

Figure 3.2 A Polynomial Based Predistorter

The following analysis derives the optimal polynomial predistorter coefficients for maximum cancellation of 3rd order intermodulation products. The predistorter circuit in figure 3.2 may be analysed as follows:

for a general two tone test with equal tone amplitudes, the input signal, $x(t)$, is given by:

$$x(t)=A\{\cos(\omega_1 t+\phi_1)+\cos(\omega_2 t+\phi_2)\} \tag{3.8}$$

This signal is applied to the amplitude and phase controllers of the predistorter giving:

$$x_2(t)=k_1 A\{\cos(\omega_1 t+\phi_1+\Phi)+\cos(\omega_2 t+\phi_2+\Phi)\} \tag{3.9}$$

where k_1 is the resultant gain of the buffer amplifier, as controlled by the variable attenuator and Φ is the phase-shift introduced by the variable phase-shift element. Note that the above assumes that the phase-shift element introduces a constant phase-shift with frequency (since the two tones are assumed to be different in frequency).

This is then applied to the cubic predistorter element, giving

$$x_3=k_2 k_1^3 A^3\{\cos(\omega_1 t+\phi_1+\Phi)+\cos(\omega_2 t+\phi_2+\Phi)\}^3 \tag{3.10}$$

By expansion this becomes

$$x_3(t)=\left(\frac{k_2 k_1{}^3 A^3}{4}\right)\left\{\begin{array}{l} 9\cos(\omega_1 t+\phi_1+\Phi)+9\cos(\omega_2 t+\phi_2+\Phi) \\ +3\cos((2\omega_1-\omega_2)t+2\phi_1-\phi_2+\Phi)+3\cos((2\omega_2-\omega_1)t+2\phi_2-\phi_1+\Phi) \\ +\cos(3\omega_1 t+3\phi_1+3\Phi)+\cos(3\omega_1 t+3\phi_1+3\Phi) \\ +3\cos((2\omega_1+\omega_2)t+2\phi_1+\phi_2+3\Phi)+3\cos((2\omega_2+\omega_1)t+2\phi_2+\phi_1+3\Phi) \end{array}\right\} \tag{3.11}$$

Assuming that the final amplifier in the secondary path has a bandpass characteristic, (which will typically be the case for a practical amplifier) the out-of-band frequencies will be removed. The secondary path input to the combiner is therefore:

$$x_4(t)=\left(\frac{3k_2 k_1{}^3 A^3}{4}\right)\left\{\begin{array}{l} 3\cos(\omega_1 t+\phi_1+\Phi)+3\cos(\omega_2 t+\phi_2+\Phi) \\ +\cos((2\omega_1-\omega_2)t+2\phi_1-\phi_2+\Phi)+\cos((2\omega_2-\omega_1)t+2\phi_2-\phi_1+\Phi) \end{array}\right\} \tag{3.12}$$

The predistorter output is given by

$$y(t)=x(t)+x_4(t) \tag{3.13}$$

Thus

$$y(t) = A\{\cos(\omega_1 t + \phi_1) + \cos(\omega_2 t + \phi_2)\}$$

$$+\left(\frac{3k_2 k_1^3 A^3}{4}\right)\begin{Bmatrix} 3\cos(\omega_1 t + \phi_1 + \Phi) + 3\cos(\omega_2 t + \phi_2 + \Phi) \\ +\cos((2\omega_1 - \omega_2)t + 2\phi_1 - \phi_2 + \Phi) + \cos((2\omega_2 - \omega_1)t + 2\phi_2 - \phi_1 + \Phi) \end{Bmatrix}$$

(3.14)

Equation 3.14 consists of the original two tones plus the intermodulation products generated by the cubic non-linearity.

The amplifier transfer function is given by

$$z(t) = a_1 y(t) + a_3 (y(t))^3 \tag{3.15}$$

By substitution of equation 3.14 this gives

$$Z(t) = a_1 \begin{bmatrix} A\{\cos(\omega_1 t + \phi_1) + \cos(\omega_2 t + \phi_2)\} \\ +\left(\frac{3k_2 k_1^3 A^3}{4}\right)\begin{Bmatrix} 3\cos(\omega_1 t + \phi_1 + \Phi) + 3\cos(\omega_2 t + \phi_2 + \Phi) \\ +\cos((2\omega_1 - \omega_2)t + 2\phi_1 - \phi_2 + \Phi) + \cos((2\omega_2 - \omega_1)t + 2\phi_2 - \phi_1 + \Phi) \end{Bmatrix} \end{bmatrix}$$

$$+a_3 \begin{bmatrix} A\{\cos(\omega_1 t + \phi_1) + \cos(\omega_2 t + \phi_2)\} \\ +\left(\frac{3k_2 k_1^3 A^3}{4}\right)\begin{Bmatrix} 3\cos(\omega_1 t + \phi_1 + \Phi) + 3\cos(\omega_2 t + \phi_2 + \Phi) \\ +\cos((2\omega_1 - \omega_2)t + 2\phi_1 - \phi_2 + \Phi) + \cos((2\omega_2 - \omega_1)t + 2\phi_2 - \phi_1 + \Phi) \end{Bmatrix} \end{bmatrix}^3$$

(3.16)

This expansion yields a large number of terms; the only terms necessary in this analysis are those at frequencies coincident with the fundamental and third-order IMP. The fundamental frequency components are given by

$$z_{FD}(t)=\left(a_1A+\frac{9a_3A^3}{4}+\frac{945a_3k_1{}^6k_2^2A^7}{32}\right)\{\cos(\omega_1t+\phi_1)+\cos(\omega_2t+\phi_2)\}$$

$$+\left(\frac{9a_1k_1^3k_2A^3}{4}+\frac{45a_3k_1^3k_2A^5}{4}+\frac{5103a_3k_1^9k_2^3A^9}{128}\right)\{\cos(\omega_1t+\phi_1+\Phi)+\cos(\omega_2t+\phi_2+\Phi)\}$$

$$+\left(\frac{45a_3k_1^3k_2A^5}{8}\right)\{\cos(\omega_1t+\phi_1-\Phi)+\cos(\omega_2t+\phi_2-\Phi)\}$$

$$+\left(\frac{945a_3k_1^6k_2^2A^7}{64}\right)\{\cos(\omega_1t+\phi_1+2\Phi)+\cos(\omega_2t+\phi_2+2\Phi)\}$$

(3.17)

Whilst the third-order IMP components are given by

$$Z_{IM}(t)=\left(\frac{3a_3A^3}{4}+\frac{567a_3k_1^6k_2^2A^7}{32}\right)\{\cos((2\omega_1-\omega_2)t+2\phi_1-\phi_2)+\cos((2\omega_2-\omega_1)t+2\phi_2-\phi_1)\}$$
$$+\left(\frac{3a_1k_1^3k_2A^3}{4}+\frac{45a_3k_1^3k_2A^5}{8}+\frac{1701a_3k_1^9k_2^3A^9}{64}\right)\times$$
$$\{\cos((2\omega_1-\omega_2)t+2\phi_1-\phi_2+\Phi)+3\cos((2\omega_2-\omega_1)t+2\phi_2-\phi_1+\Phi)\}$$
$$+\left(\frac{45a_3k_1^3k_2A^5}{16}\right)\{\cos((2\omega_1-\omega_2)t+2\phi_1-\phi_2-\Phi)+3\cos((2\omega_2-\omega_1)t+2\phi_2-\phi_1-\Phi)\}$$
$$+\left(\frac{567a_3k_1^6k_2^2A^7}{64}\right)\{\cos((2\omega_1-\omega_2)t+2\phi_1-\phi_2+2\Phi)+3\cos((2\omega_2-\omega_1)t+2\phi_2-\phi_1+2\Phi)\}$$

(3.18)

Each vector can be seen to consist of four components; it is the resultant of these components which is required to be found. If a single frequency is considered (either one of the fundamental tones or a single IMP), the resultant phasor may be seen to be of the form

$$R\cos(\theta+\gamma)=E\cos(\theta)+F\cos(\theta+\Phi)+G\cos(\theta-\Phi)+H\cos(\theta+2\Phi) \qquad (3.19)$$

By expansion and simplification this becomes

$$R\cos(\theta+\gamma)=\cos\theta\{E+(F+G)\cos\Phi+H\cos2\Phi\}-\sin\theta\{(F-G)\sin\Phi+H\sin2\Phi\}$$

(3.20)

In order to find the conditions for optimum cancellation the amplitude of the phasor, R, is required,

$$R^2 = [E + (F+G)\cos\Phi + H\cos 2\Phi]^2 + [(F-G)\sin\Phi + H\sin 2\Phi]^2 \tag{3.21}$$

By expansion this becomes

$$R^2 = E^2 + F^2 + G^2 + H^2 + 2\cos\Phi\{EF + EG + FH\} + 2\cos 2\Phi\{FG + EH\} + 2\cos 3\Phi\{GH\} \tag{3.22}$$

In this manner, the magnitudes of the fundamental and the third-order IMP components may be found. The fundamental amplitude may be found from:

$$R_F^2 = \left(\frac{A^2}{16384}\right)\left[\begin{array}{l} \left\{\begin{array}{l} 16384a_1^2 + 73728a_1a_3A^2 + 82944a_3^2A^4 + 82944a_1^2k_1^6k_2^2A^4 \\ +1797120a_1a_3k_1^6k_2^2A^6 + 4769280a_3^2k_1^6k_2^2A^8 + 2939328a_1a_3k_1^{12}k_2^4A^{10} \\ +32557140a_3^2k_1^{12}k_2^4A^{12}26040609a_3^2k_1^{18}k_2^6A^{16} \end{array}\right\} \\ +\left\{\begin{array}{l} 73728a_1^2k_1^3k_2A^2 + 718848a_1a_3k_1^3k_2A^4 + 1244160a_3^2k_1^3k_2A^6 \\ +4572288a_1a_3k_1^9k_2^3A^8 + 24712128a_3^2k_1^9k_2^3A^{10} \\ +57868020a_3^2k_1^{15}k_2^5A^{14} \end{array}\right\}\cos\Phi \\ +\left\{898560a_1a_3k_1^6k_2^2A^6 + 3162240a_3^2k_1^6k_2^2A^8 + 21636720a_3^2k_1^{12}k_2^4A^{12}\right\}\cos 2\Phi \\ +\left\{2721600a_3^2k_1^9k_2^3A^{10}\right\}\cos 3\Phi \end{array}\right] \tag{3.23}$$

In a similar manner, the magnitude of the resultant third-order IMP may be found from:

$$R_{IM}^2 = \left(\frac{9A^6}{4096}\right)\left[\begin{array}{l} \left\{\begin{array}{l} 256a_3^2 + 256a_1^2k_1^6k_2^2 + 3840a_1a_3k_1^6k_2^2A^2 + 30096a_3^2k_1^6k_2^2A^4 \\ +18144a_1a_3k_1^{12}k_2^4A^6 + 314685a_3^2k_1^{12}k_2^4A^8 + 321489a_3^2k_1^{18}k_2^6A^{12} \end{array}\right\} \\ +\left\{\begin{array}{l} 512a_1a_3k_1^3k_2 + 5760a_3^2k_1^3k_2A^2 + 18144a_1a_3k_1^9k_2^3A^4 \\ +199584a_3^2k_1^9k_2^3A^6 + 642978a_3^2k_1^{15}k_2^5A^{10} \end{array}\right\}\cos\Phi \\ +\left\{1920a_1a_3k_1^6k_2^2A^2 + 20448a_3^2k_1^6k_2^2A^4 + 210924a_3^2k_1^{12}k_2^4A^8\right\}\cos 2\Phi \\ +\left\{22680a_3^2k_1^9k_2^3A^6\right\}\cos 3\Phi \end{array}\right] \tag{3.24}$$

In order to illustrate the analysis, an amplifier operating at a certain compression point is used (usually the 1dB power compression point). The peak-input amplitude is set to come to this level. The degree of compression is given by the ratio of the actual gain and the linear gain (from equation 3.15)

$$C_p = \frac{a_1 + a_3\hat{A}^2}{a_1} \tag{3.25}$$

To calculate the third-order IMP level at this input amplitude, equation 3.25 may be used. The ratio of the third-order IMP amplitude to the fundamental is given by

$$R_{IMP} = 20\log_{10}\left|\frac{\left(\frac{3a_3A^3}{4}\right)}{a_1A + \left(\frac{9a_3A^3}{4}\right)}\right| \tag{3.26}$$

Or

$$R_{IMP} = 20\log_{10}\left|\frac{3a_3A^2}{4a_1 + 9a_3A^2}\right| \tag{3.27}$$

This is similar to the expression derived for feedforward cancellation [6]. Consider the following parameters:

$$A = 0.5,\ a_1 = 1,\ a_3 = -0.2057$$

From equation 3.27 the resultant R_{IMP} is

$$R_{IMP} = -27.21 \quad \text{dBc} \tag{3.28}$$

The analysis shows that the result is very similar to feedforward cancellation, as seen from the predistorter output, there is vector addition of the cubic distortion produced by the amplifier of the linear part of the predistortion signal, and the linearly amplified cubic predistortion signal. Vector addition is, in general only applicable to linear systems since it is a form of superposition. However, it is also applicable here, as the output of the amplifier is being considered, rather than the input.

3.3.1 Gain and Phase Error

Using equation 3.24, it is possible to show how imperfect gain and phase balance settings for the predistorter will influence the degree of IMD cancellation that can be achieved. This is shown in figure 3.3 for various gain (amplitude) and phase errors prior to the cubic non-linearity (in the lower path of the predistorter) and in figure 3.4 if these errors occur after the cubic non-linearity. In practice, errors will occur at both points, and so the result will be a combination of the two effects.

These results are similar (in form) to that derived in [7] for feedforward correction and examination of them leads to the conclusion that an ideal predistorter would require an extremely high degree of gain and phase matching accuracy in order to achieve a high level of IMD reduction. This is unlikely to be necessary for conventional diode or transistor-based predistorters, as their characteristics are not sufficiently accurate to provide high levels (>30dB) of IMD removal; hence a more relaxed gain and phase matching specification can be used. It is also worth noting the differences between figure 3.3 and figure 3.4; with the former representing gain and phase errors prior to the cubic non-linearity and the latter errors occurring after it. It is evident that the system is, in general more tolerant of amplitude errors occurring after the cubic non-linearity and roughly equally tolerant of phase errors in either position, for a given amount of IMD suppression (clearly, at a given gain error there is a difference in response to phase errors in the two positions, however this arises from the difference in achievable IMD suppression and not a fundamental difference between the two positions). A comparison of figure 3.3 and figure 3.4 leads to the conclusion that any gain adjustment used in setting up the predistorter (whether manually or automatically controlled) would be better placed after the cubic non-linearity, as this position is most tolerant of gain errors. An alternative (and equivalent) position would be prior to or following the time delay element in the upper path of figure 3.2. This may prove to be a more convenient location in some applications, due to the lower signal levels present in the lower path, and the consequent need to minimise losses whenever possible.

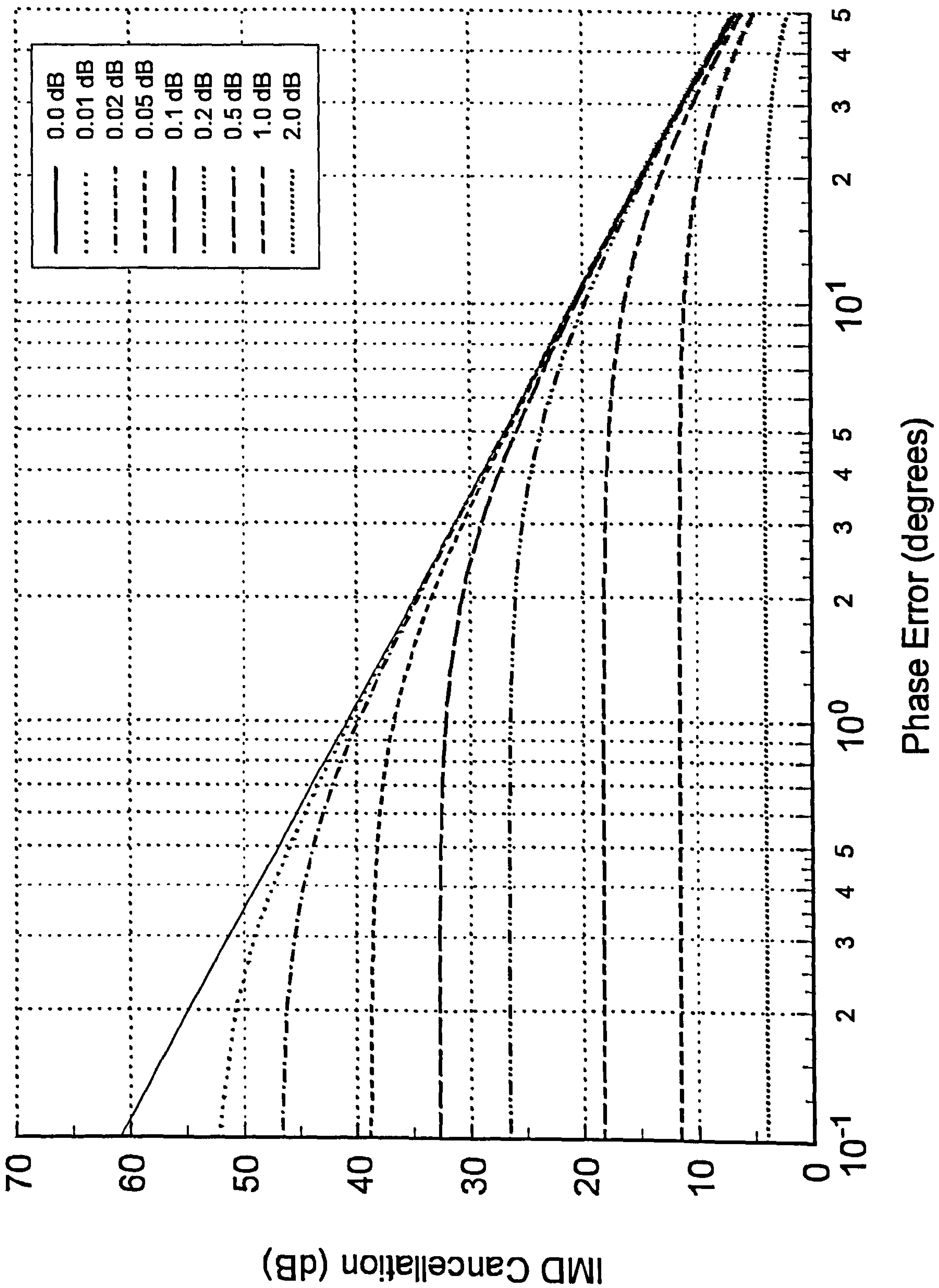

Figure 3.3 IMD cancellation which can be achieved for various values of gain and phase error, prior to the cubic non-linearity

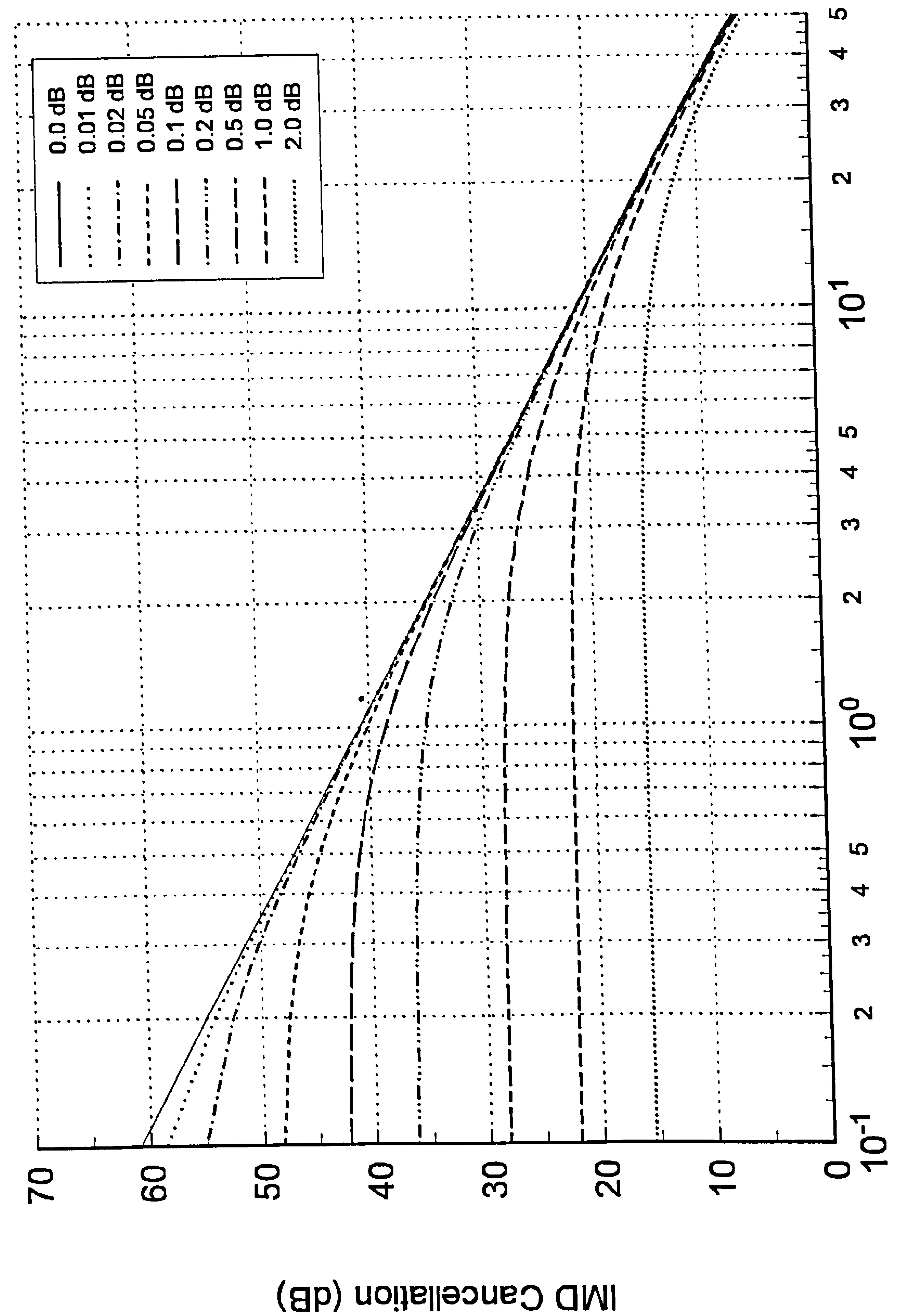

Figure 3.4 IMD cancellation which can be achieved for various values of gain and phase error, after the cubic non-linearity

The matching requirements in a predistortion system appear to be marginally less critical than those required for an equivalent degree of IMD removal in a feedforward system. For example, a 15dB improvement in IMD performance may be achieved with amplitude and phase accuracy, following the cubic non-linearity, in the region of 2dB and 10 degrees respectively. This degree of matching is relatively easy to achieve, without recourse to any form of automatic control system. In order to achieve the same level of performance from a feedforward system, the amplitude match would need to be improved to 1dB; alternatively, 1.5dB would suffice with a corresponding improvement in phase accuracy (to around 5 degrees). The above results have assumed a purely AM-AM characteristic for the RF power amplifier and this will not be the case in practice, as AM-PM distortion will have a significant effect. This will further degrade the likely performance level of a simple predistorter of the type considered here. It is, however, possible to use two such predistorters in a quadrature arrangement in order to correct for both AM-AM and AM-PM distortion and this arrangement should offer an improved overall performance. It should, of course, be remembered that the improvements achieved in figure 3.3 and figure 3.4 are additional to the raw IMD performance of the RF power amplifier. So, for example, if the uncompensated main amplifier had a third-order IMD ratio of 30dB and the predistortion system had a gain error of 0.5dB and a phase error of less than 1 degrees (following the cubic non-linearity), the overall third-order IMD performance of the complete system would be better than about -55dBc.

In practice, however, the performance of a purely third-order predistorter will be limited by the fifth-order IMD level of the RF power amplifier. Therefore there is little point in suppressing the third-order products to a level much below the level of the fifth-order IMD.

3.3.2 Intermodulation Distortion Performance

In addition, by using equation 3.27, it is possible to relate the compression point of an amplifier with a purely third-order characteristic, to the resultant level of third-order IMD it will produce when driven at that level. This is shown in table 3.1, for various values of compression point.

Compression level (dB)	*Third-order IMD ratio (dBc)*
0.1	47.3
0.2	41.3
0.5	33.3
1.0	27.2
1.2	25.6
1.5	23.7
2.0	21.2
2.5	19.3
3.0	17.7
4.0	15.4

Table 3.1 Relationship between the compression point and IMD ratio of an amplifier with a purely third-order non-linearity

This data allows a simple AM/AM amplifier model to be used with a known level of IMD ratio and compression level. It also gives an approximate idea of the compression level at which an amplifier is operating directly from its two-tone test result. The accuracy of this approximation will decrease, however, as the non-linearity level in the amplifier increases. It is not, therefore, appropriate for highly non-linear amplifiers.

The operation of the RF predistortion system described above is best illustrated by the use of a two-tone input signal. If such a signal is applied to the amplifier specified in the previous section, with the input level chosen to drive the amplifier approximately 1dB into compression, the resulting spectrum is as shown in figure 3.5.

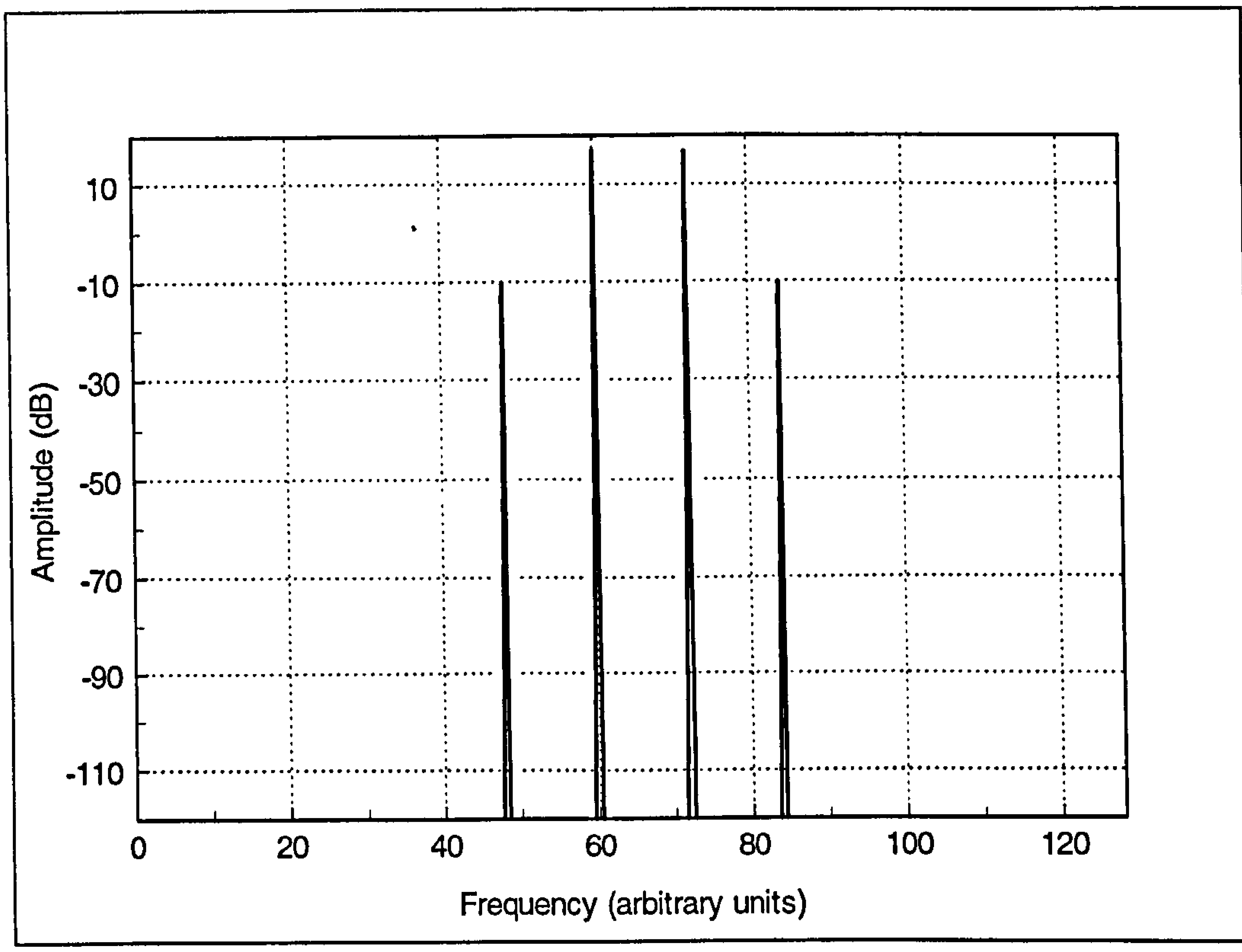

Figure 3.5 Output Spectrum for an amplifier with linear gain and a third-order non-linearity, operating with a peak output power at its 1dB compression point.

Note that only third-order products are present (in addition to the original input tones), since a purely third-order non-linearity has been assumed for the amplifier. If the form of predistorter described in the previous section, is used prior to the PA whose two-tone response is shown in figure 3.5, the resulting output spectrum is that shown in figure 3.6 (assuming an ideal setting for the variable attenuator, and perfect subtraction at the summing junction).

Note that higher-order intermodulation products have now been introduced, despite the power amplifier (PA) having only a third-order non-linearity. Note also that these products are well above the level of the third-order products (now eliminated) and hence provide the fundamental limit to the achievable linearity. These additional products are present at a relatively high level, despite purely third-order characteristics being assumed for the PA and the non-linear predistorting element.

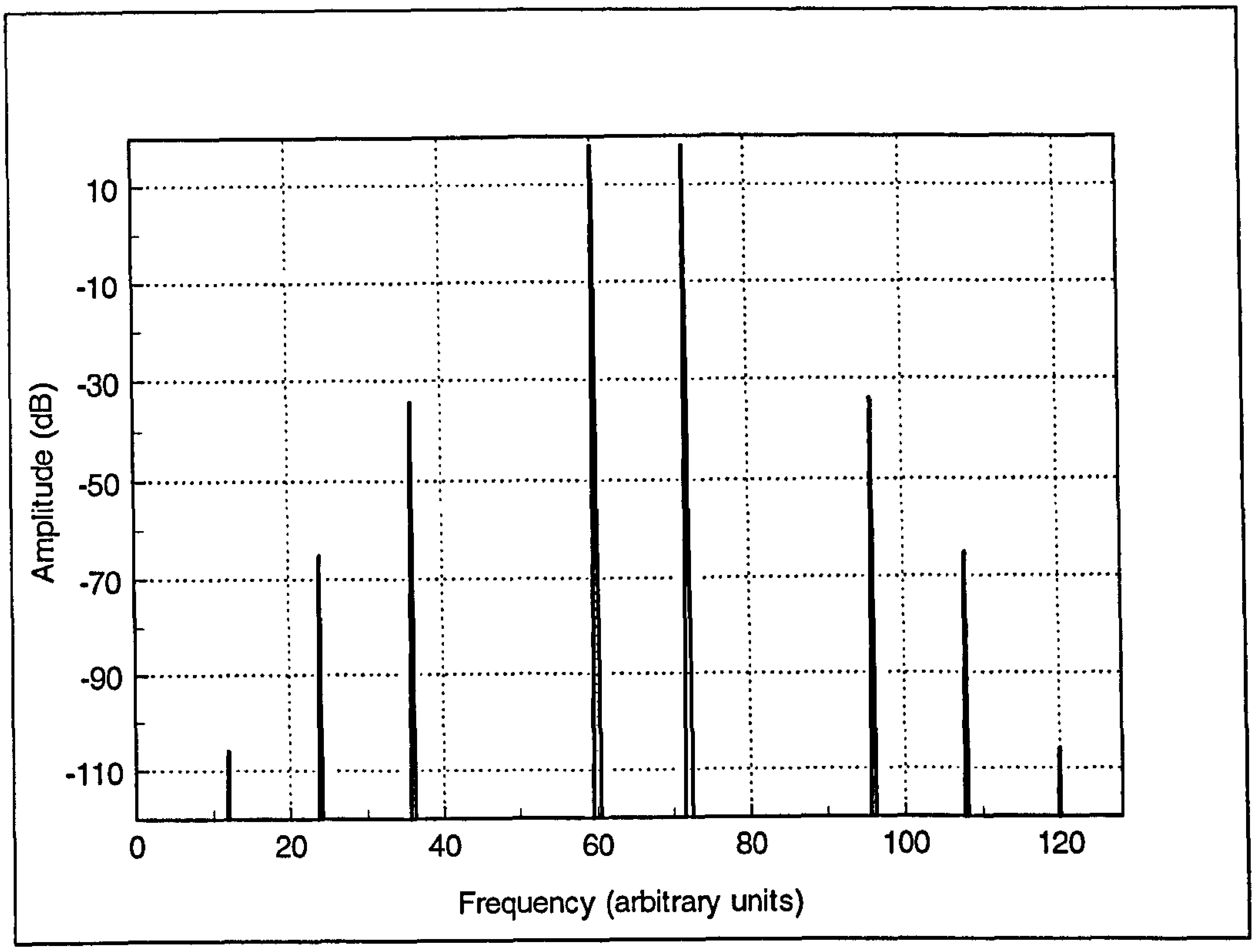

Figure 3.6 Output spectrum for an amplifier with linear gain and third-order non-linearity, when preceded by an ideal third-order predistorter.

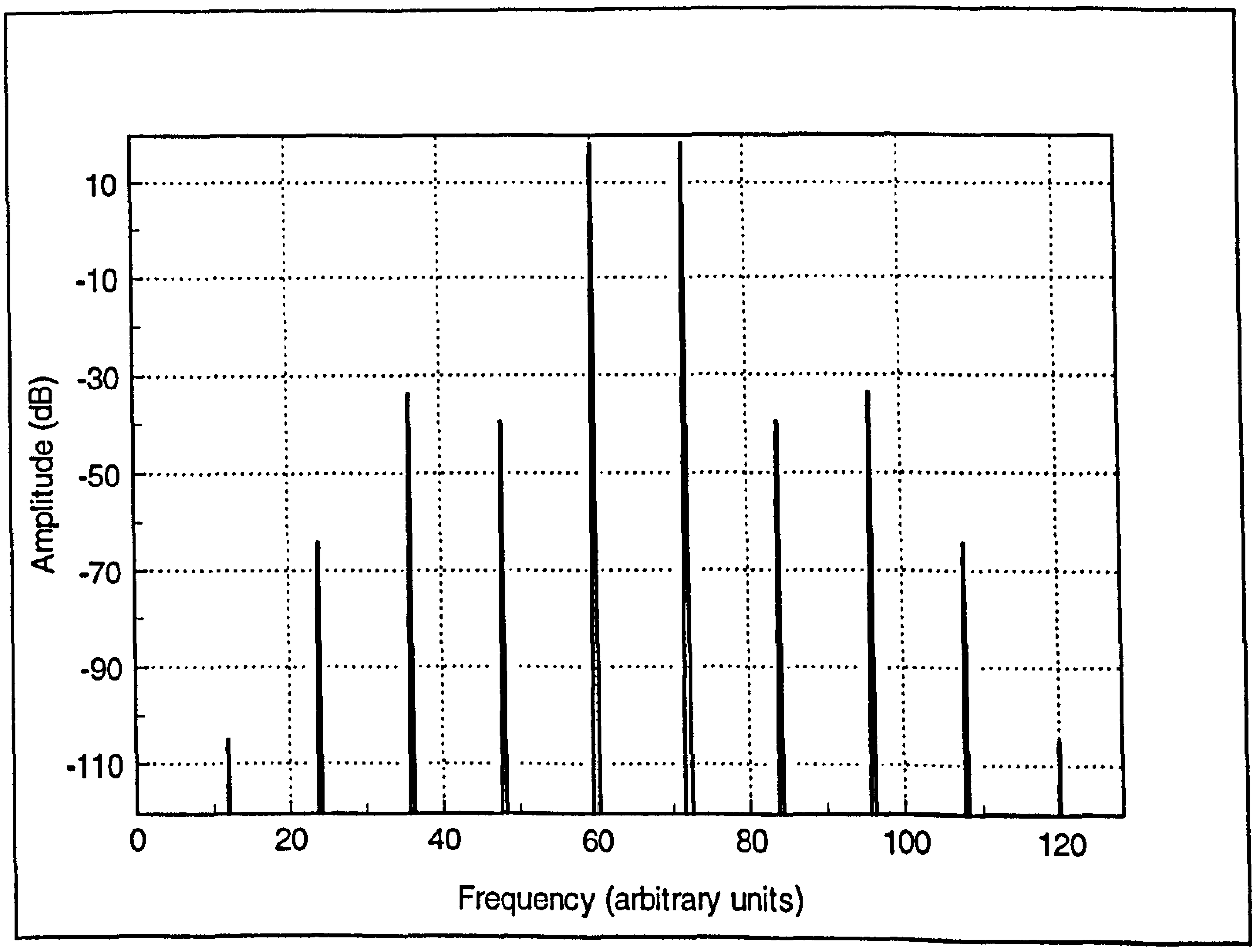

Figure 3.7 Output spectrum for the amplifier when preceded by a third-order predistorter with a 0.1dB gain error in its lower path.

Finally, if it is assumed that the variable attenuator is incorrectly set (with the phase-shifter correctly set), and it introduces a 0.1dB error into the lower path of the predistorter (prior to the cubic non-linearity), the resulting spectrum is shown in figure 3.7. The third-order IMD cancellation is degraded, as would be expected, but the level of the third-order products is still below that of the fifth-order products (which have not altered, despite the 0.1dB error). This demonstrates the relatively generous margin for error which this type of predistorter allows, whilst still achieving a high level of performance (around 40dBc overall).

3.4 Summary

This chapter has shown the mathematical requirements for ideal predistortion linearisation of an amplifier. A new analysis for a cubic predistortion system has been introduced and results are presented which show that for perfect cancellation of the third order IMP's the gain and phase matching requirements are extremely tight. Gain and phase matching requirements are presented for errors that occur both before and after the predistorting element. From these results it is proposed that the gain and phase adjustment be placed after the cubic non-linearity to allow optimum cancellation of IMP's even in the presence of gain and phase errors. This chapter has shown that theoretically it is possible to achieve very high levels of cancellation of the third order IMP's using this technique.

References

[1] P. Bura, D. Gelerman and P. Ntake, Linear Solid State Power Amplifier for 64 QAM Radios, ICC, volume 1, pages 1 – 5, June 1987.

[2] G. P. Locatelli and L. Ricco and M. Nannicini, Microwave Linear Power Amplifier with Micromodule Technology, IEEE Transactions on the International Communications Conference, volume 1, pages 11 – 16, June 1987.

[3] P. B. Kenington and K. J. Parsons and D. W. Bennett, Broadband linearisation of high-efficiency power amplifiers, Proceedings of the Third International Mobile Satellite Conference, pages 59 – 64, June 1993.

[4] K. P. Parsons, Highly Efficient Broadband Linearisation of R. F. Amplifiers for Satellite and Cellular Applications, University of Bristol, March 1996.

[5] Herbert. L. Kraus, Charles. W. Bostain and Frederick. H. Raab, Solid State Radio Engineering, John Wiley & Sons New York, 1980.

[6] P. B. Kenington, Efficiency of Feedforward Amplifiers, IEE Proceedings, volume 139, pages 591 – 593, October 1992.

[7] R. J. Wilkinson and P. B. Kenington, Specification of error amplifiers for use in feedforward transmitters, IEE Proceedings, volume 139, No. 4, pages 477 - 480, August 1992.

Chapter 4
Practical Cubic Predistorter

This Chapter describes a polynomial predistorter with a new type of polynomial generator element. This predistorter is able to provide improvements in predistortion performance over much greater bandwidths than have been achieved previously.

4 Practical Cubic Predistorter

4.1 Introduction

Polynomial predistortion [1, 2, 3] has the potential to provide useful improvements in the linearity of an R.F. power amplifier whilst not adversely affecting the efficiency of the amplifier being linearised. The importance of this fact should not be under estimated when considering the form of linearisation applied to a particular amplifier which is required to be linearised. Traditional polynomial predistortion has either been carried out at baseband using DSP techniques [4, 5] or attempts have been made to generate approximately polynomial characteristics using diodes biased near their cut off point or FET's used near their pinch off regions [6, 7, 8, 9, 10, 11, 12, 13]. DSP techniques will in general provide the greatest level of predistortion linearisation improvement but they are limited in bandwidth. DSP predistorters require very well balanced up and down conversion chains to ensure correct matching of the predistorter to the amplifier characteristic. DSP predistorters also have a substantial impact on the efficiency of the overall system due to the power requirements of the DSP. Diodes, BJT's and FET's when used to generate polynomial functions only provide the crudest approximation to the true polynomial function which is required. The main cause of this non-ideal performance is due to the method employed to generate the function. The device is always biased on the knee of the characteristic, which results in additional orders of polynomial coefficients being generated. Use of this type of polynomial generator has been confined to the early work on the identification of a truly 'generic' predistorter[1]. This chapter introduces a new type of polynomial predistortion element, which uses simple hardware to predistort an amplifier over far higher bandwidths than have been reported previously.

[1] Generic predistorters are predistorters that attempt to produce an exact inverse of the amplifier being linearised. This type of predistorter may be considered as an ideal predistorter.

4.2 A New Type of Polynomial Predistorter

4.2.1 Mixer Based Polynomial Predistorter

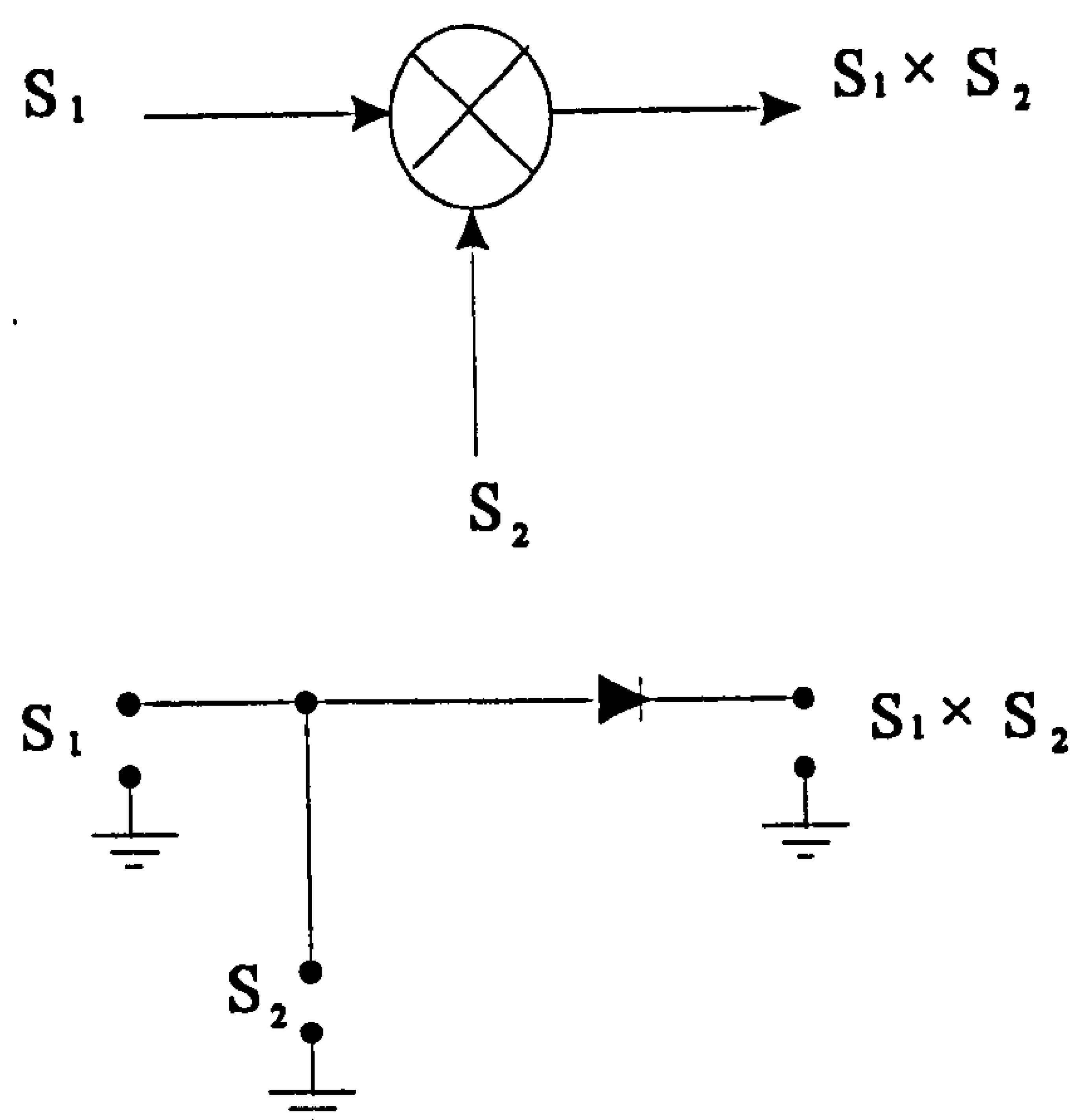

Figure 4.1 A Mixer

A mixer shown in figure 4.1 is a non-linear component that causes sum and difference frequencies of the input signals to be generated. The general form of the output verses input non-linearity can be expressed mathematically as a power series such as [14]:

$$i_0 = I_o + av_i + bv_i^2 + cv_i^3 + \ldots + nv_i^n \tag{4.1}$$

where: I_o is the DC current, av_i to nv_i^n are fundamental and 2nd, 3rd to nth order distortion terms. This equation shows that the mixer exhibits a non-linear relationship that may be exploited to generate polynomial predistorter.

4.2.2 The N^{th} Order Polynomial Predistorter

Mixers may be cascaded together to form a polynomial predistorter which is shown in figure 4.2.

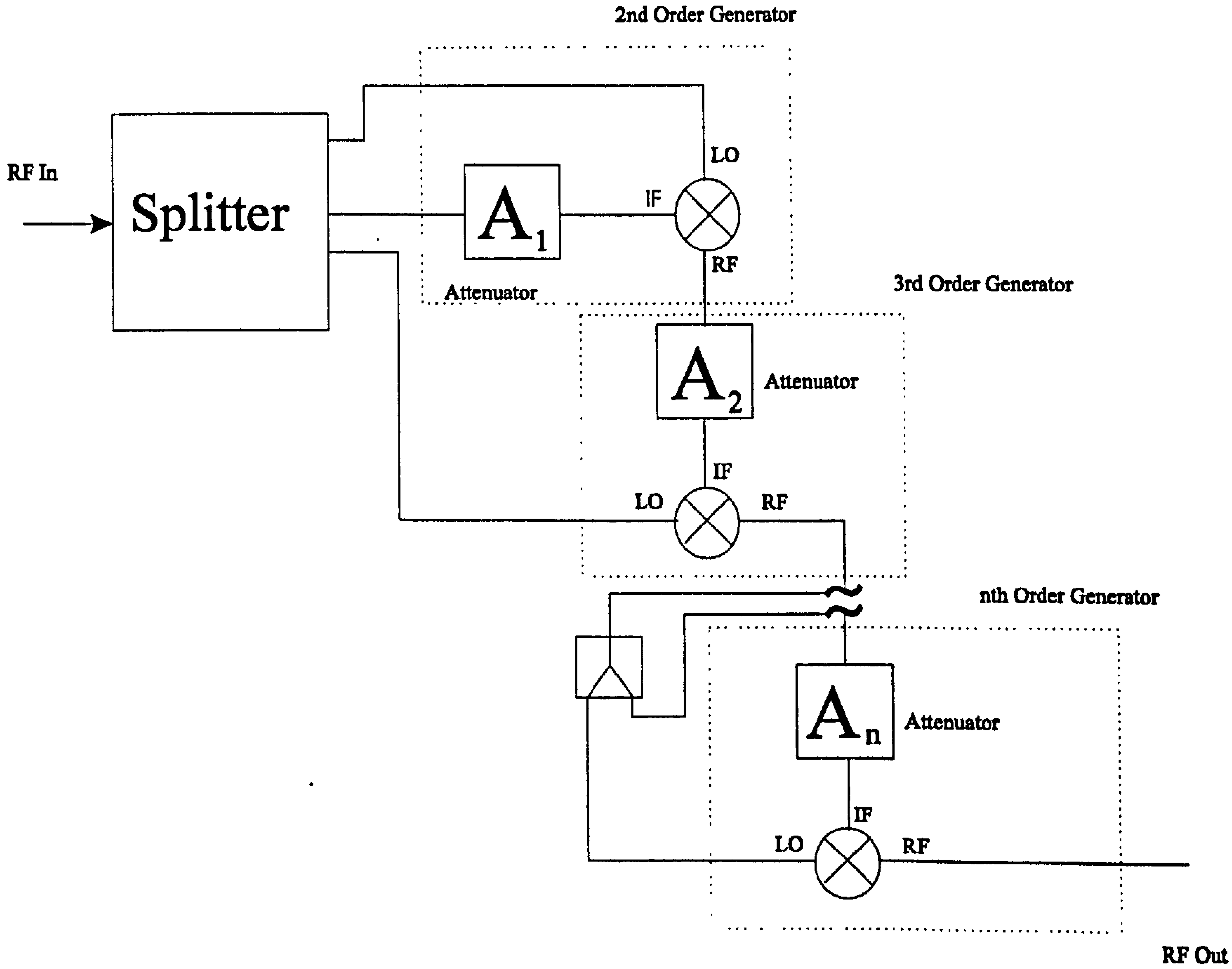

Figure 4.2 The N^{th} Order Polynomial Predistorter

This arrangement is the general design for an n^{th} order polynomial predistorter. The mixers provide a multiplication function such that the LO and IF signals are multiplied together with the resultant output appearing at the RF port. Mixers are cascaded together as shown such that a single mixer produces a second order nonlinearity, two mixers produce a cubic nonlinearity, three mixers produce a quintic nonlinearity etc. In the theory this method could be used to generate any order of nonlinearity that was required. Initially investigations have centred on producing a cubic nonlinearity. This is due to the 3^{rd} order intermodulation products being the most significant in the amplifiers requiring linearisation.

4.3 Practical Investigation of the 3rd Order Polynomial Predistorter

4.3.1 The Cubic Non Linearity

When designing a practical cubic predistorter shown in figure 4.3, it is important that the mixers are driven within their specified linear operating region. The local oscillator (LO) ports must be driven so that conversion loss is at a minimum. Conversion loss increases as drive level is reduced. Manufacturers specify mixers in terms of frequency range and LO drive level. For minimum conversion loss mixers should be driven at the specified LO level. When mixer LO drive level is above the specified level then power is wasted. If the mixer is driven above its absolute maximum power level then damage will result. The intermediate frequency (IF) port has a 1dB compression point typically 20dB below the LO drive level. In order to operate well below this point attenuation is required before each IF port. If this is not the case then highly unpredictable non-linear behavior occurs which results in poor generation of 3rd order products. This added attenuation results in a symmetrical and predictable cubic non-linearity. Values of attenuation were chosen as follows $A_1 = 40$dB and $A_2 = 20$dB. The design utilized a ZFSC 3-4 three-way power splitter and two LRMS 2D level 7 mixers from Mini Circuits [16].

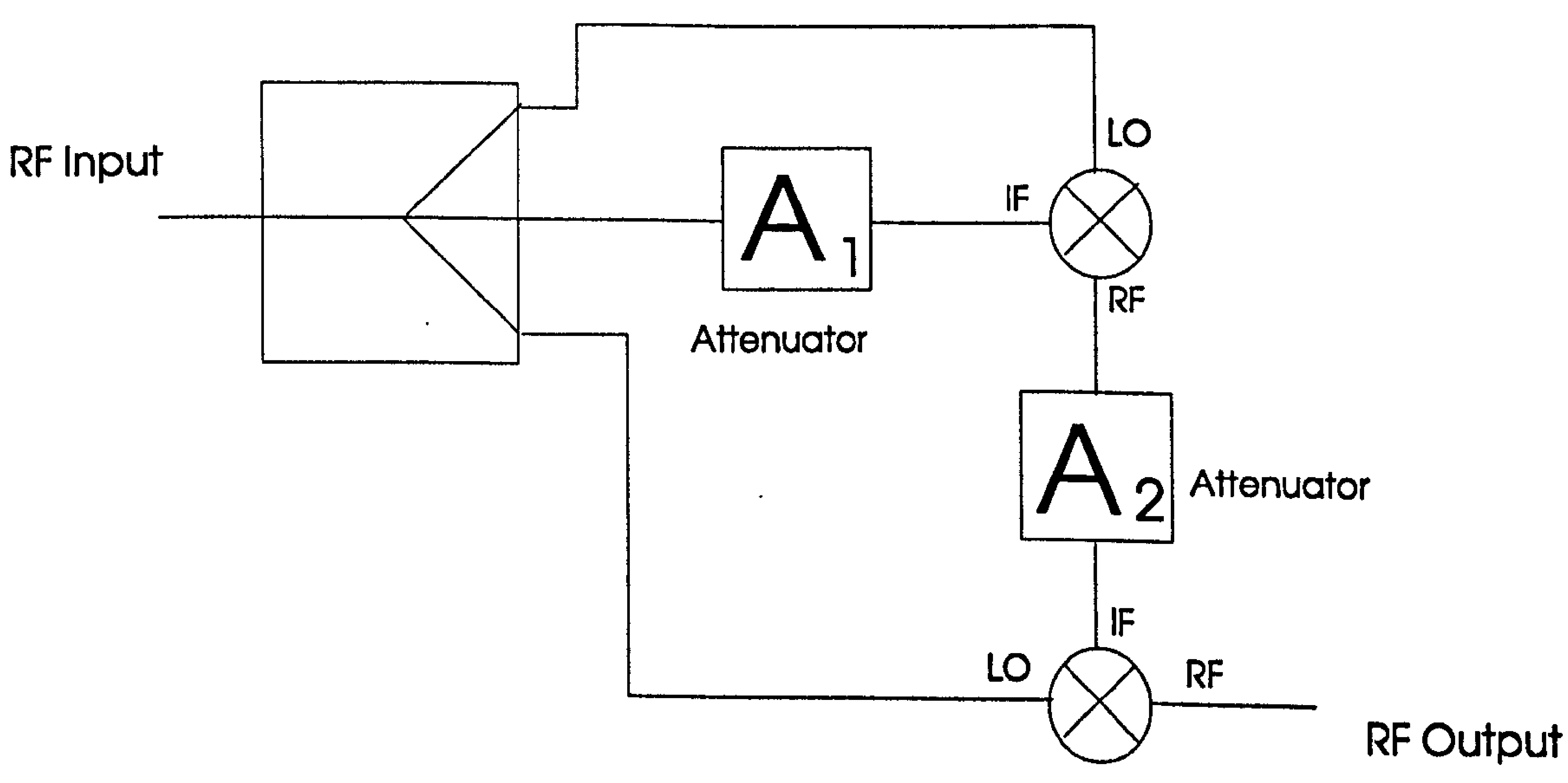

Figure 4.3 The 3rd Order Polynomial Predistorter

A two-tone test was applied to the cubic predistorter element to investigate its performance. The signal consisted of two tones with a centre frequency of 850MHz, a tone spacing of 500kHz and an input level of 9dBm. The resultant output is shown in figure 4.4.

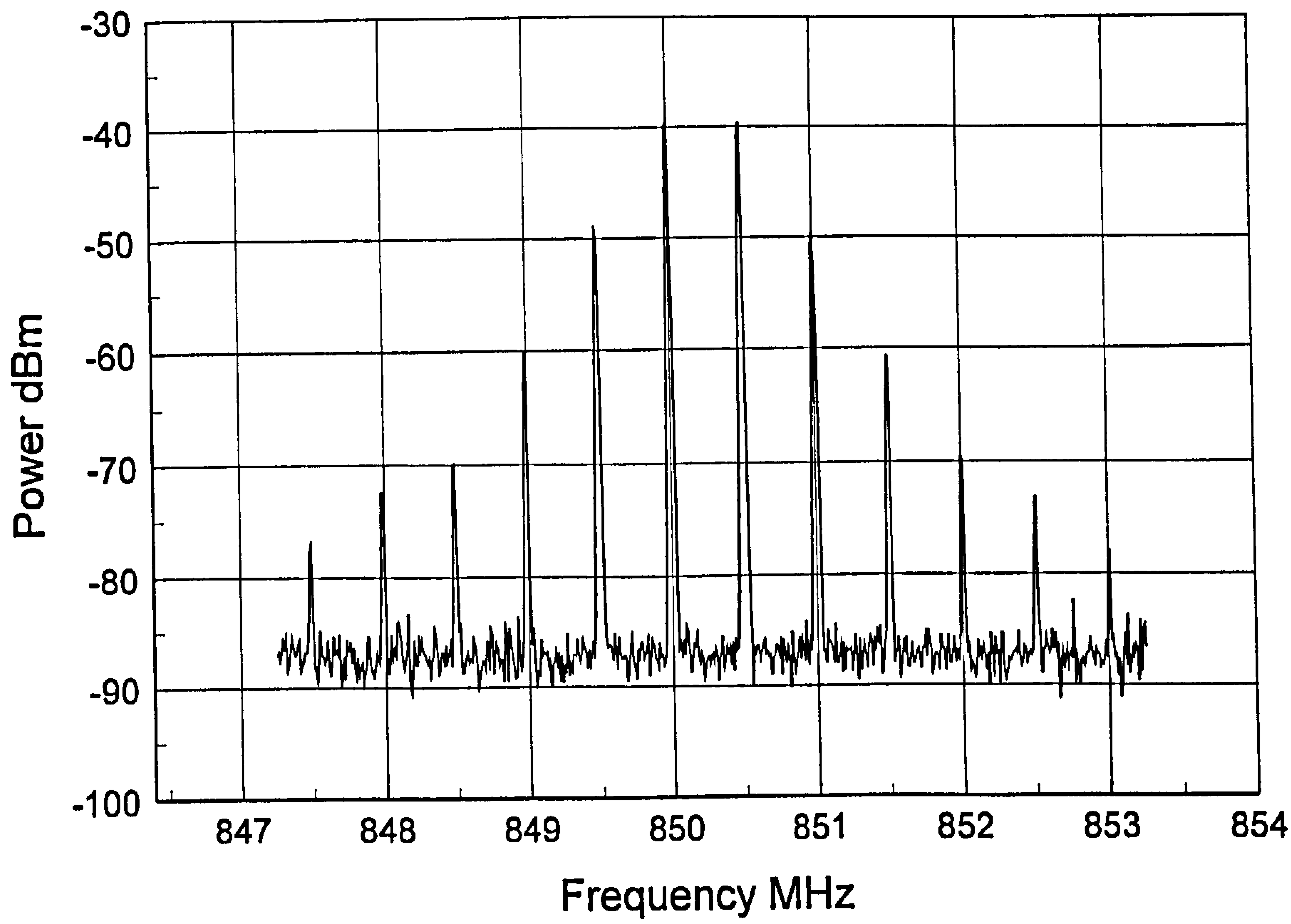

Figure 4.4 The 3rd Order Polynomial Predistorter Output

The plot clearly shows the symmetrical nature of the characteristic. It also shows that the characteristic contains orders of non-linearity greater than 3^{rd}. The 3^{rd} order terms are 10dB below the two tones that produced them. With each successive term 10dB below the preceding one. With reference to equation 3.16 a cubic non-linearity should in theory produce IMP energy up to 9^{th} order IMP this can also be seen in the plot. It may also be observed that 11^{th} order IMP energy is also present due to the non-ideal performance of the mixer elements.

4.3.2 The Predistortion System

The cubic non-linearity is only one element in an operational predistortion system. An operational predistorter consists of the fundamental path with attenuation and the predistortion path. The predistortion path consists of the cubic non-linearity, a buffer amplifier, and an adjustable attenuator and phase shifter, this topology is shown in figure 4.5

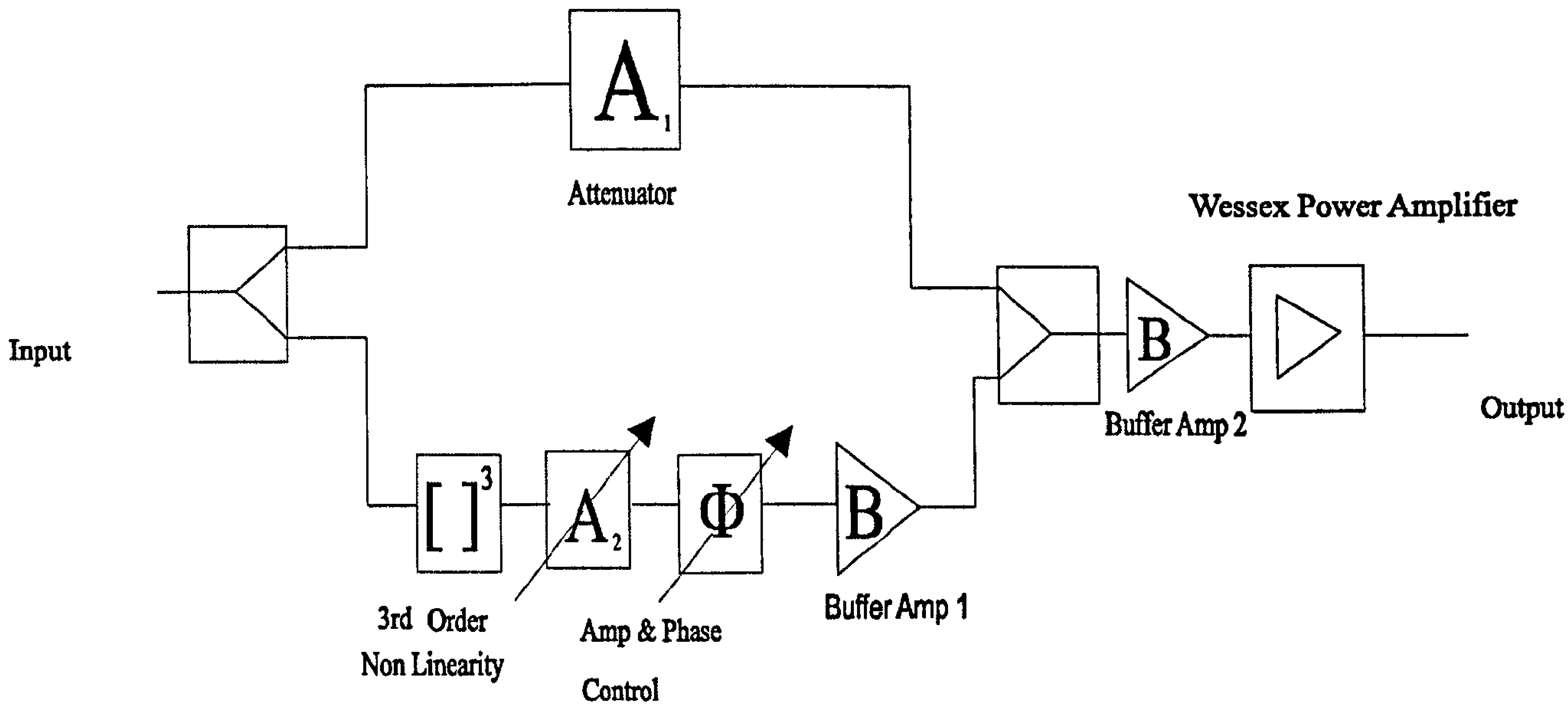

Figure 4.5 The Predistorter System

The predistorter functions as follows, the signal is split into a fundamental and predistorted path. The signal in the predistorted path first passes through the cubic non-linearity where the predistorted signal is generated. The adjustable attenuator and phase shifter provide amplitude and phase control respectively. The buffer amplifier in association with the attenuator in the forward path ensures that the 3rd order products are at the correct level below the main two tones so that 3rd order cancellation results.

The predistorter was tested with an input level of 9dBm at 850MHz and a tone spacing of 500kHz the signal being measured at the output of buffer amplifier 2 is shown in figure 4.6.

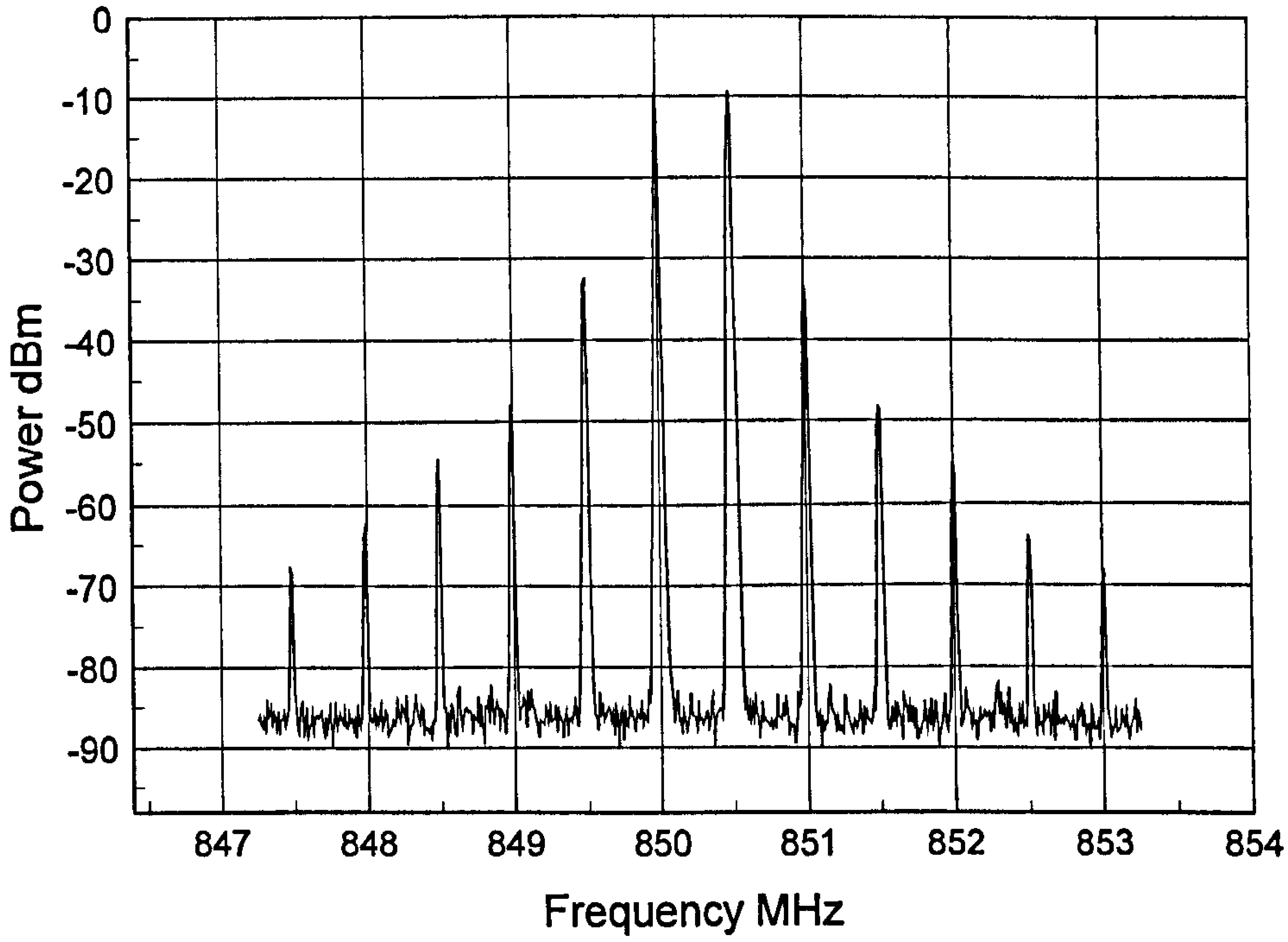

Figure 4.6 The Predistorter Output

The plot shows that the input to the Wessex amplifier is effectively a multi-tone test. With the main tones at -9dBm and the 3^{rd} order IMP's at 24dB below the main tones. An ideal predistorter would produce an output that contained main-tone and 3^{rd} order IMP's only. The plot clearly shows that additional energy is also present in the 5^{th} , 7^{th} , 9^{th} and the 11^{th} IMP's. This is due to the generation of these additional orders by the cubic nonlinearity. These additional IMP's will result in sub optimal cancellation of the 3^{rd} order IMP's.

The difference between the main-tone and 3rd order IMP energy is important in achieving the best possible cancellation. The ratio of main-tone to 3^{rd} order IMP energy at the predistorter output needs to be the same as the ratio of main-tone to third order IMP energy which is given by the Wessex amplifier two tone test in figure 4.7.

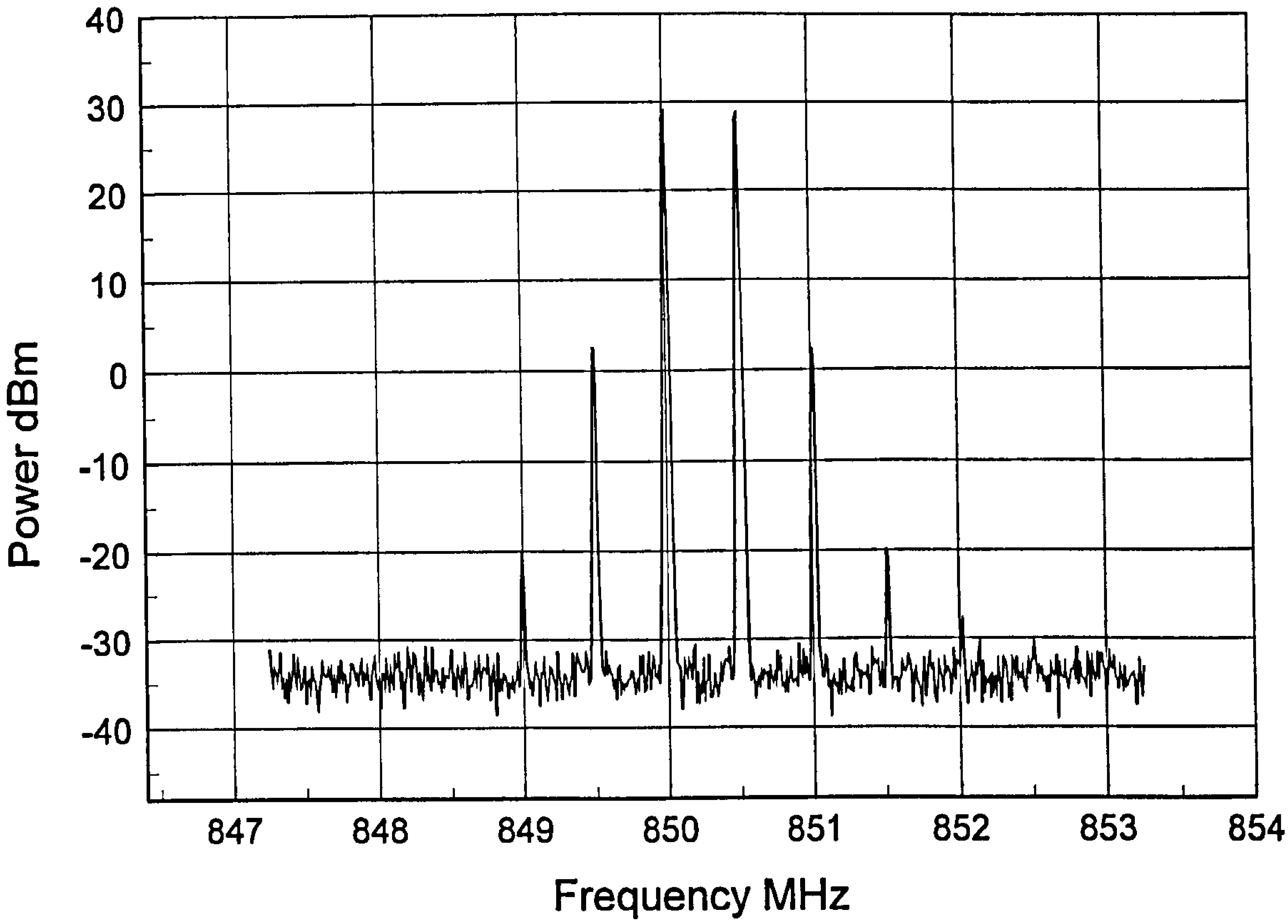

Figure 4.7 Plot of Two Tone Test at 850MHz

The plot shows that the 3rd order IMP's are 26dB below the main-tone energy i.e. -26dBc. So the predistorter needs to have a characteristic which has 3rd order IMP's at -26dBc. Applying this signal to the Wessex amplifier results in the predistorted output shown in figure 4.8.

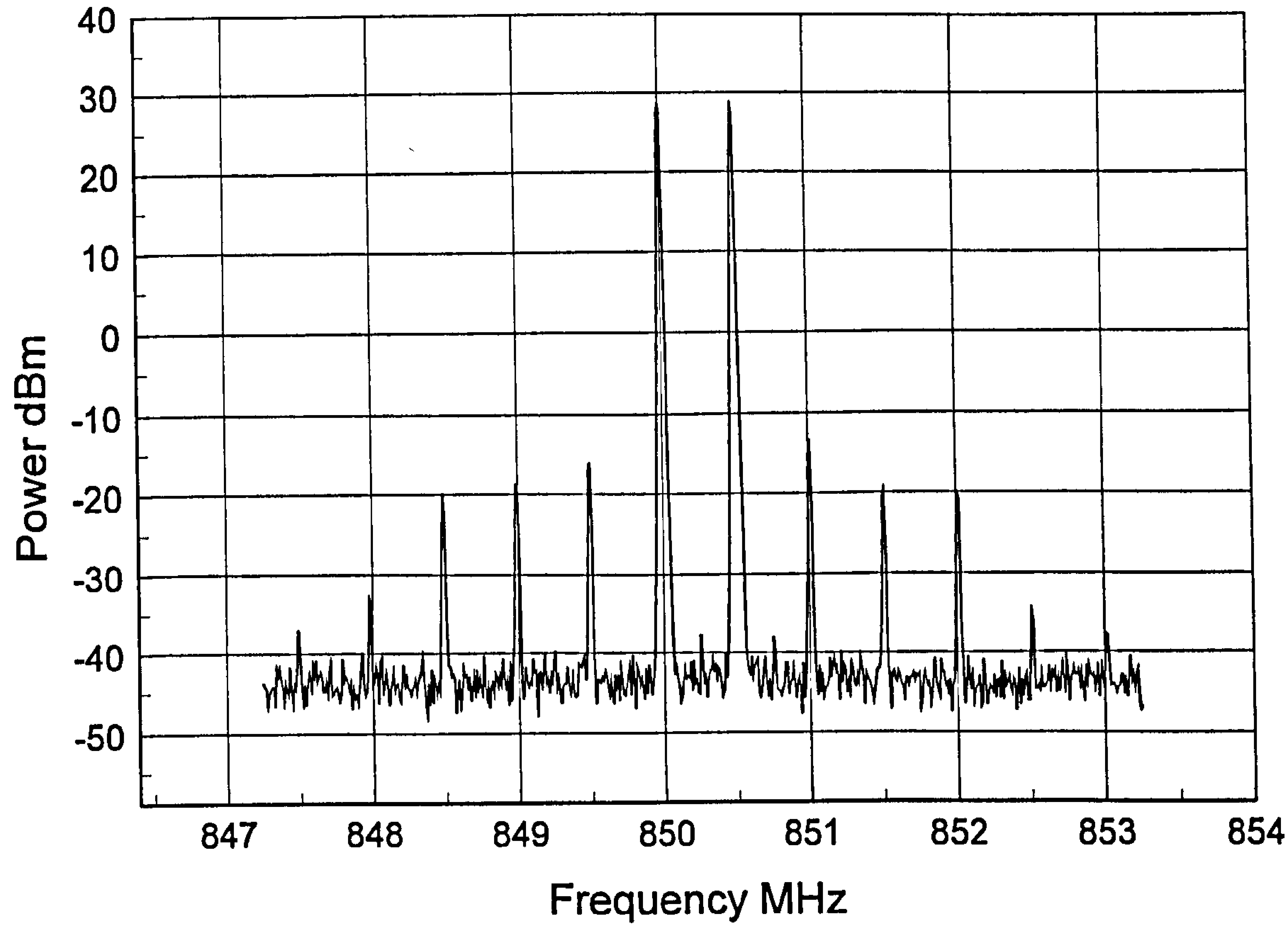

Figure 4.8 Plot of Predistorted Amplifier Output at 850MHz

The plot shows that the 3rd order IMP's are now at -44dBc. This is an 18dB improvement in performance over the two-tone test. The 5th and 9th order IMP's have increased. This is due to the generation of these terms by the cubic non-linearity. The IMD specification of an amplifier is usually quoted for the 3rd order IMP but can be quoted for other orders, especially if another order of IMP is greater than the 3^{rd} order products. In this case it can be seen that although the other orders have increased, the 3^{rd} order IMP's are still the largest.

The Wessex amplifier under test had a frequency range of 100MHz to 1GHz. There is no theoretical reason that the predistorter should not work over this range. Therefore linearisation tests were carried out at 100MHz, 220MHz, 500MHz and 1GHz. The two tone test results and the resultant Wessex amplifier predistorted outputs are shown in figures 4.9, 4.10, 4.11, 4.12, 4.13, 4.14, 4.15 and 4.16 respectively.

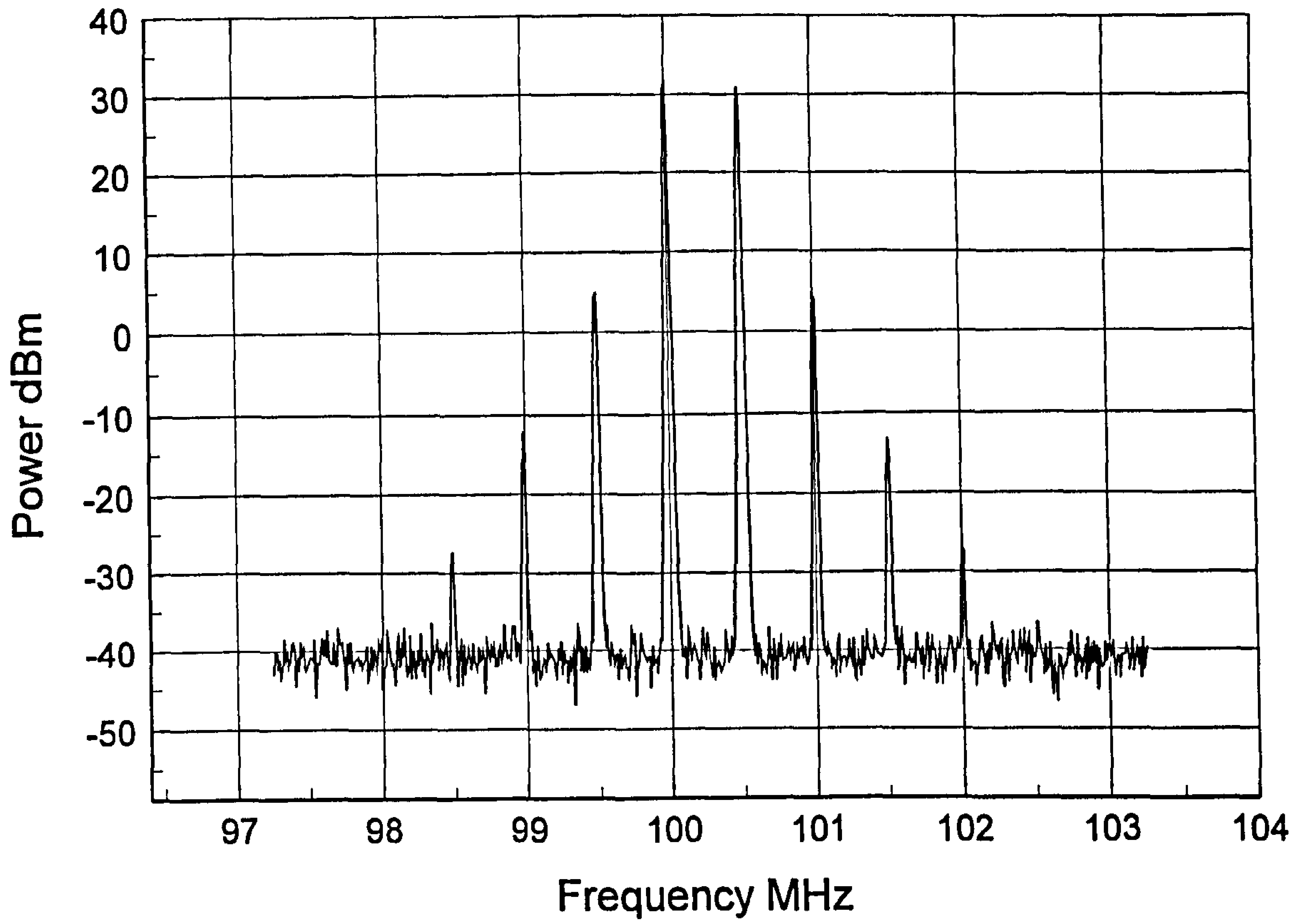

Figure 4.9 Plot of Two Tone Test at 100MHz

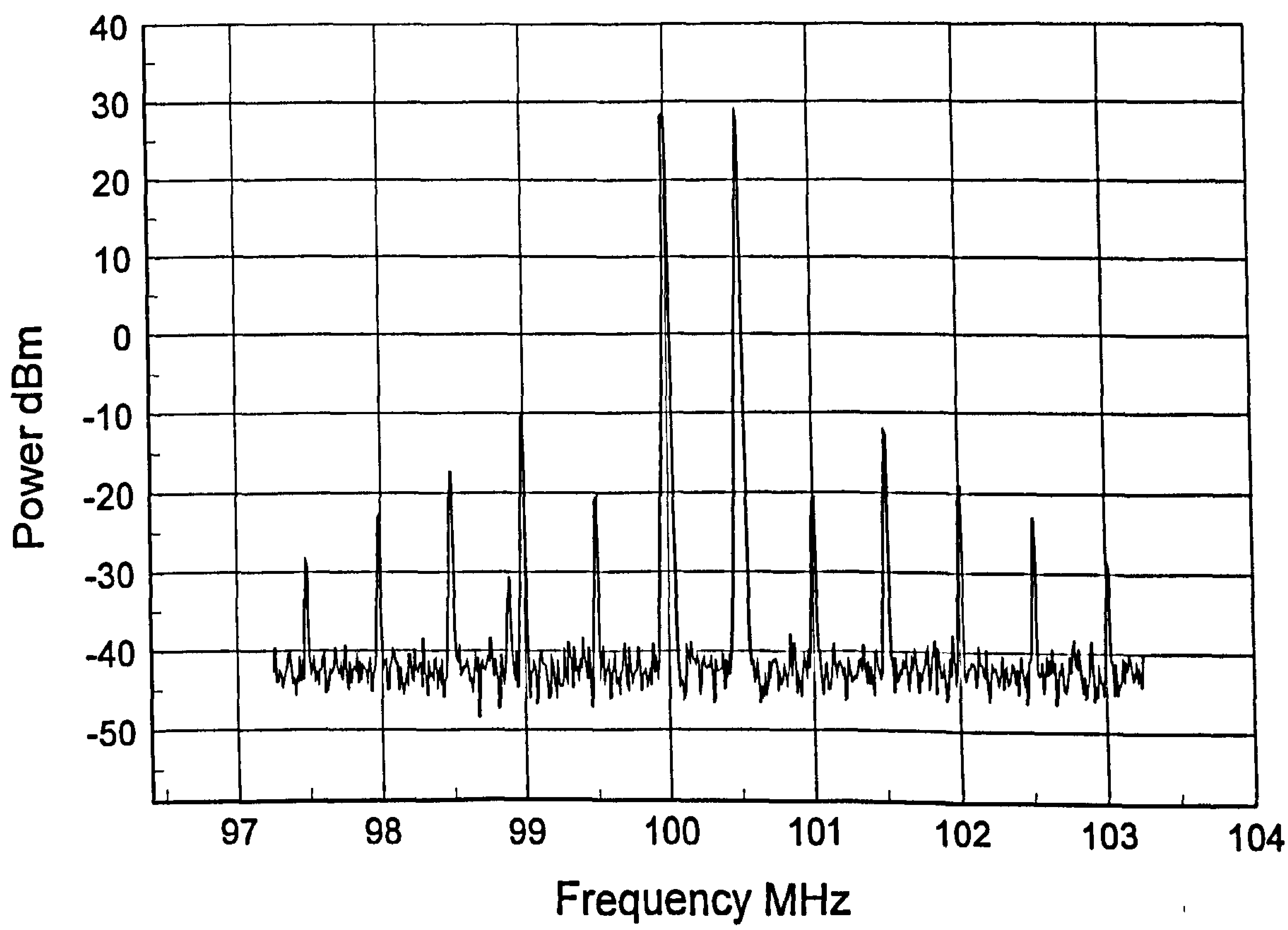

Figure 4.10 Plot of Predistorted Amplifier Output at 100MHz

The two tone test and the predistorted output at 100MHz shown in figures 4.9 and 4.10 show that the 3rd order products are cancelled by 25dB to -49dBc. The plots also show that the 5th order products however are not affected so the overall specification of the amplifier is improved by 25dB.

At 220MHz shown in figures 4.11 and 4.12, the 3rd order products are cancelled by 14dB to -46dBc, the 5th order products however have increased by approximately 8dB, so the overall specification has improved by 12dB.

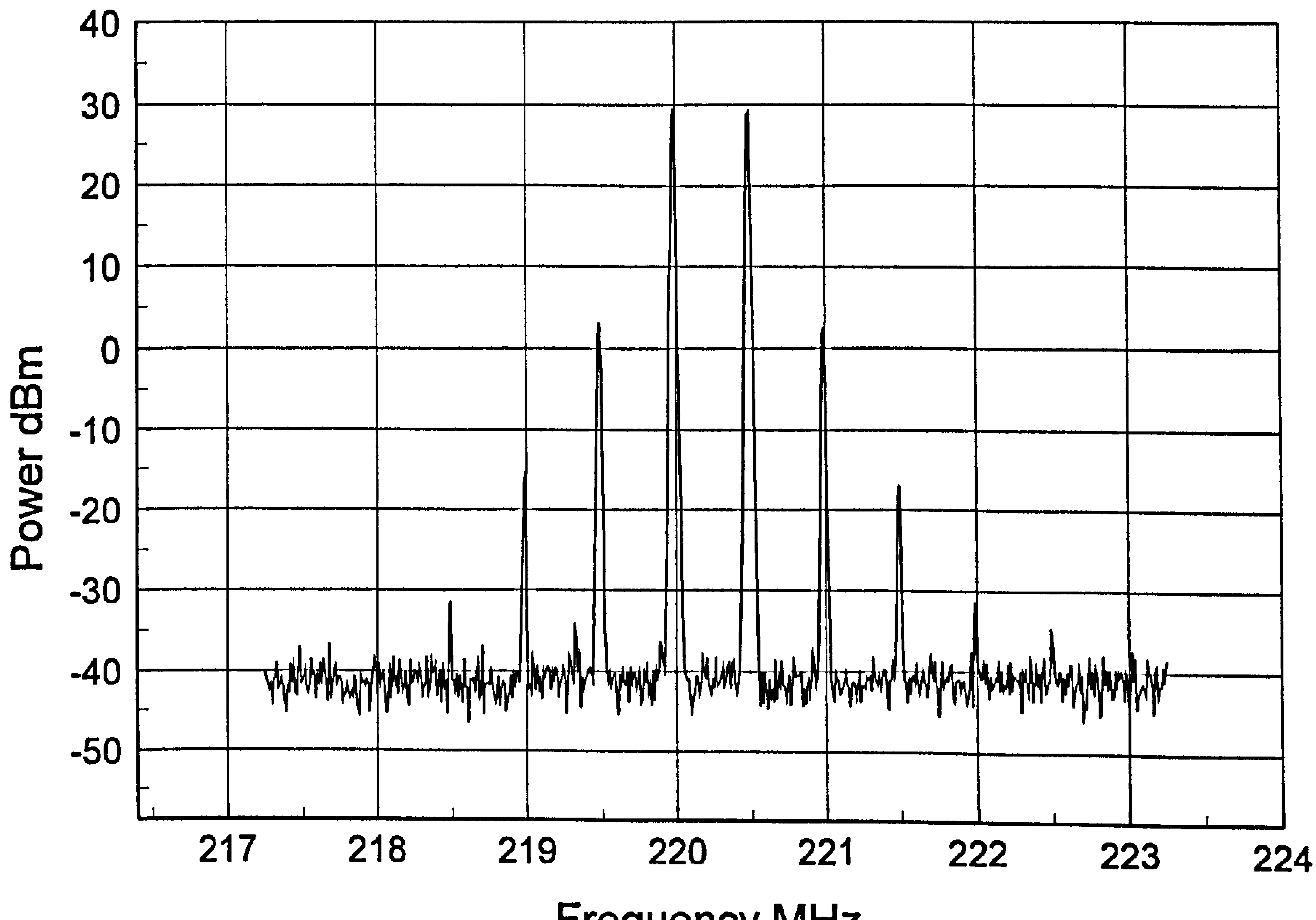

Figure 4.11 Plots of Two Tone Test at 220MHz

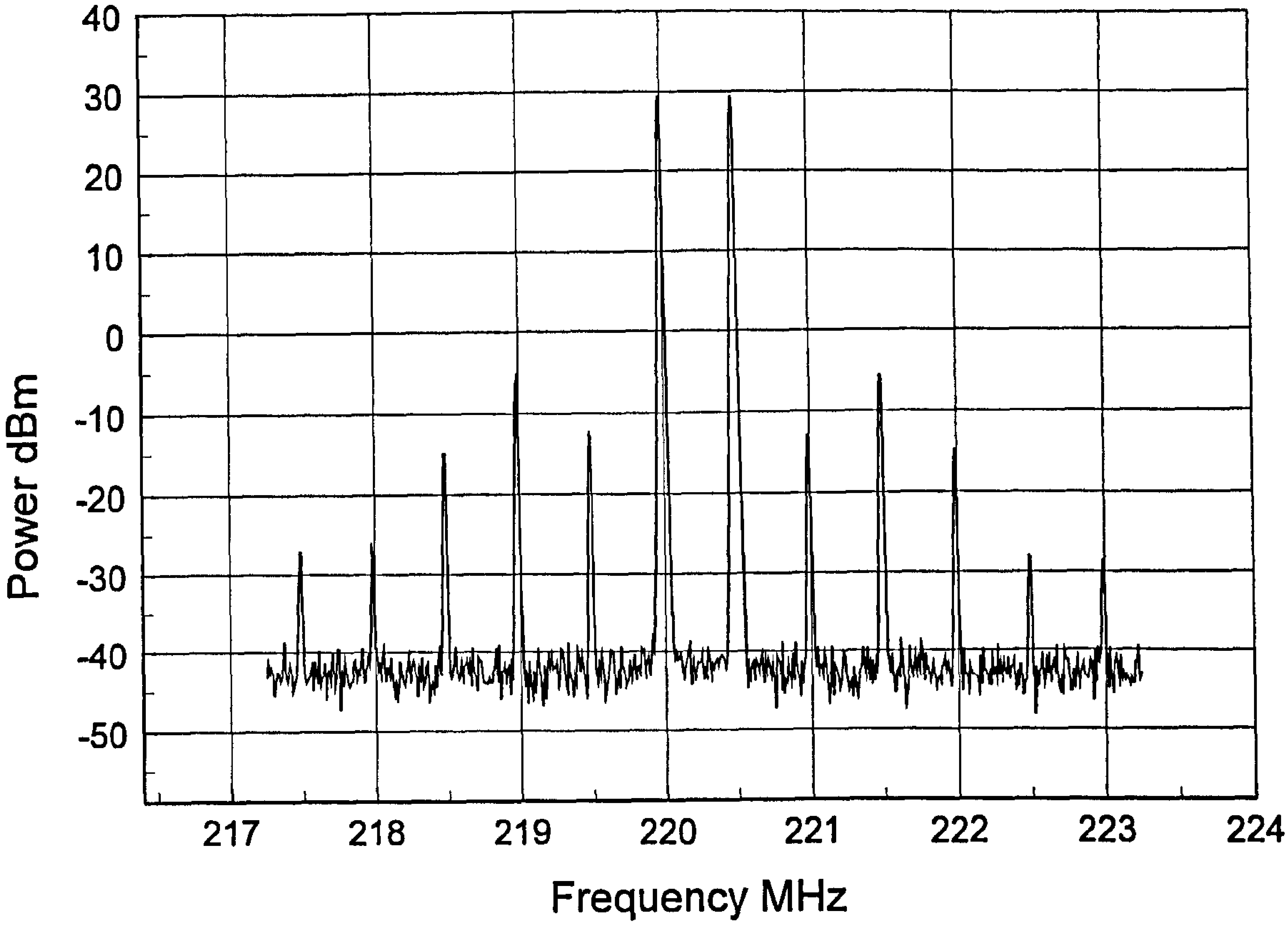

Figure 4.12 Plot of Predistorted Amplifier Output at 220MHz

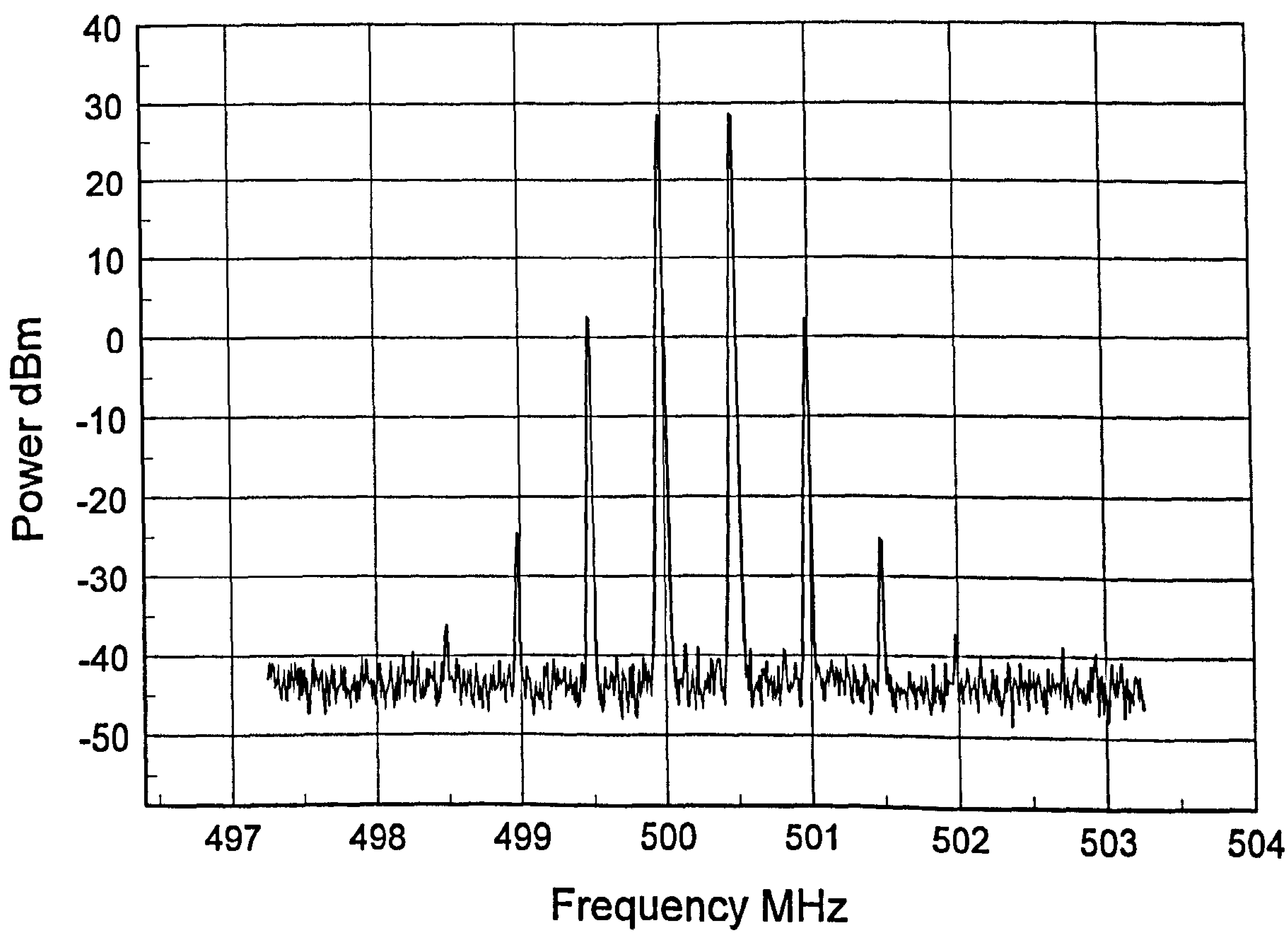

Figure 4.13 Plot of Two Tone Test at 500MHz

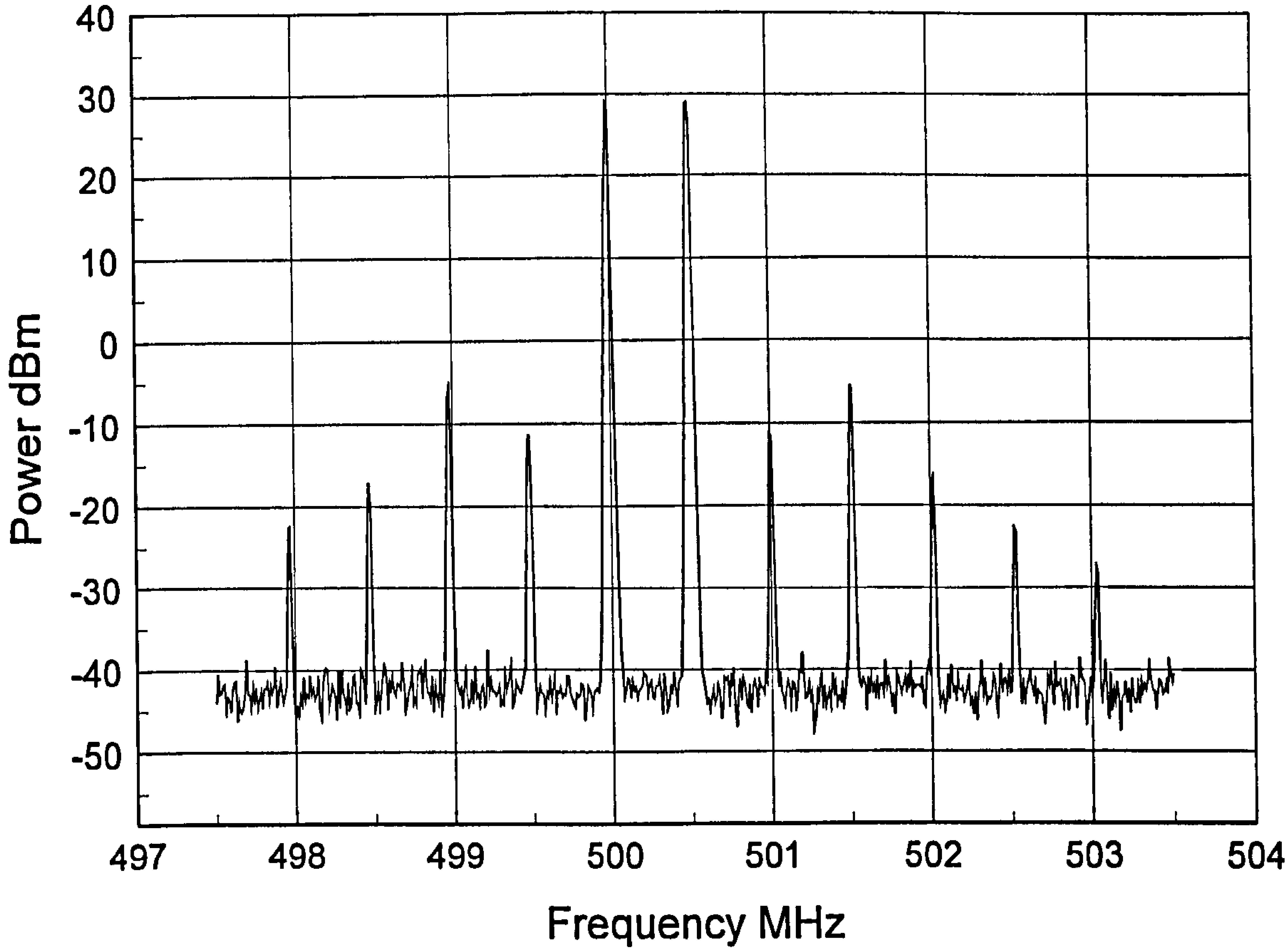

Figure 4.14 Plot of Predistorted Amplifier Output at 500MHz

At 500MHz figures 4.13 and 4.14 show that the 3[rd] order products are cancelled by 13dB to -42dBc, the 5th order products however have increased by 20dB, so the overall specification has improved by approximately 8dB.

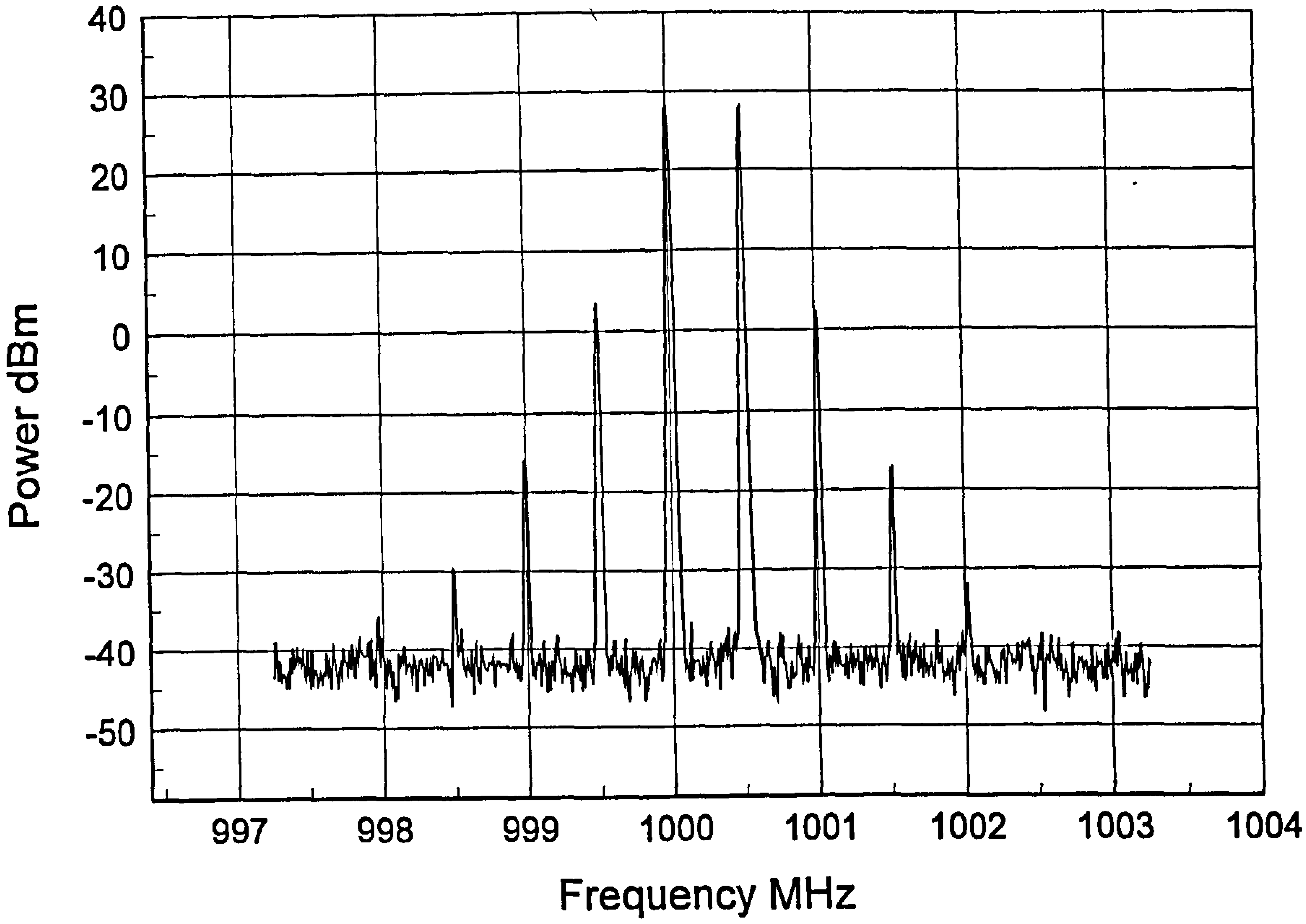

Figure 4.15 Plots of Two Tone Test at 1GHz

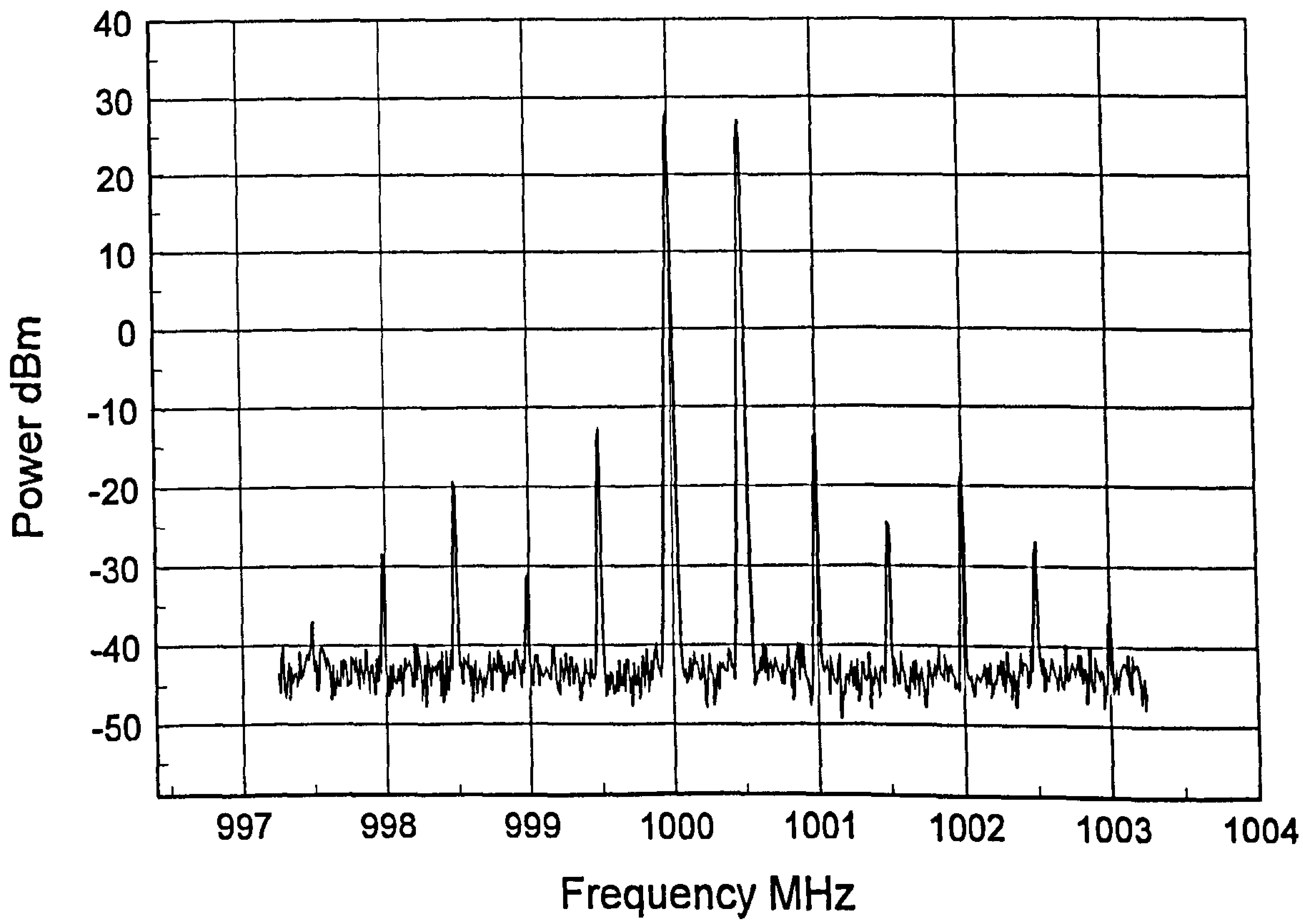

Figure 4.16 Plot of Predistorted Amplifier Output at 1GHz

At 1GHz figures 4.15 and 4.16 show that the 3rd order products are cancelled by 14dB, the 5th order products have reduced by 10dB. So the overall specification in this case is defined by the reduction in 3rd order products. This gives an overall improvement in the specification of 15dB.

All the above results were achieved with a tone spacing of 500kHz. This represents a medium bandwidth system. For the technique to operate in broadband applications linearisation must be achieved with tone spacings greater than 1MHz.

4.3.3 Electrical Delay

Initial investigations with tone spacings of 1MHz or greater gave unsatisfactory results with little or no improvement in linearity. In order to investigate the reason for this loss of performance the phase shift across the linearised amplifier was measured. The set up in figure 4.17 was used.

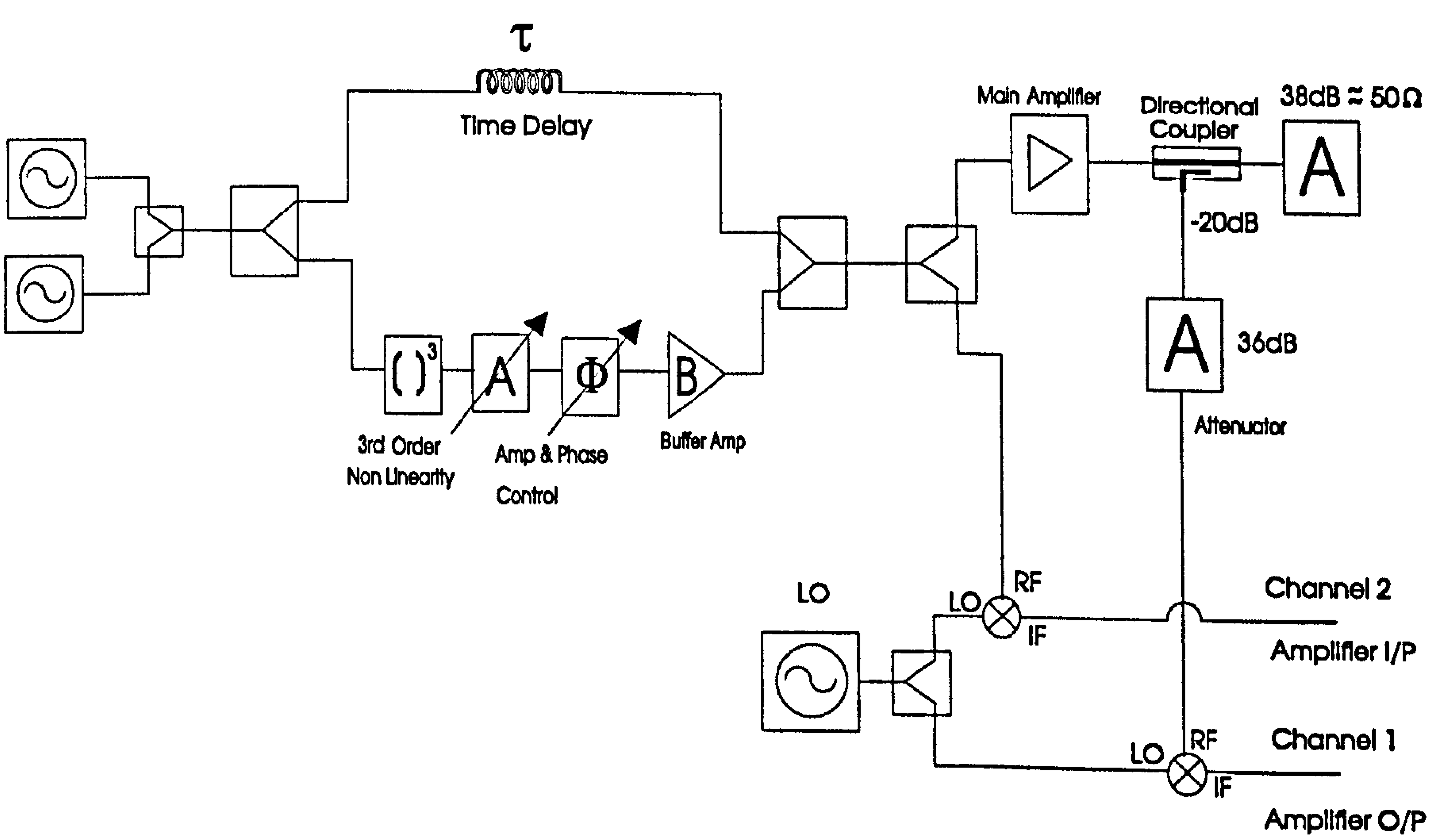

Figure 4.17 IMD Phase Measurement Set Up

The 89440A vector signal analyser was used to measure the phase difference between the main tones and the 3rd order IMP's at the amplifier output at tone spacings of 25kHz, 500kHz and 3MHz. The results of these measurements are shown in table 4.1.

Tone Spacing	Main Tone Phase Difference	3rd Order IMP Phase Difference
25 kHz	0^{o}	4^{o}
500 kHz	2^{o}	17^{o}
3 MHz	8^{o}	50^{o}

Table 4.1 Table of IMD Phase Measurement Results

The results show that as the tone spacing is increased the phase difference increased between both the main tones and the 3rd order IMP's. The phase differences for the 3MHz case when there is no cancellation are considerably higher than the phase differences for 25kHz and 500kHz spacing when acceptable cancellation results are achieved. This implies that there must be a difference in electrical delay between the fundamental and the predistorted paths. This is effectively a feedforward type effect in which the components need to arrive in the correct phase relationship in order for the system to operate correctly. Therefore the electrical delay was adjusted in the forward path to give 3rd IMP cancellation at 3MHz tone spacing. The electrical delay was then measured in both paths. The actual delay was the same in each path and equal to 16ns. To test how this modification had altered performance narrow-band and wide band measurements were taken, these are discussed in the next section.

4.3.4 Narrowband and Wideband Measurements

Measurements of achievable cancellation were taken, for the narrowband case for tone spacings of 50kHz and 100kHz and wideband measurements were taken for 30MHz and 60MHz tone spacing. All measurements were taken at a centre frequency of 900MHz. The two-tone tests and the cancellation results are shown in figures 4.18, 4.19, 4.20, 4.21, 4.22, 4.23, 4.24, and 4.25.

Comparing figures 4.18 and 4.19 it can be seen that the 3^{rd} order products are cancelled to -56dBc an improvement of 38dB over the uncorrected case. Unfortunately the 5^{th} order products are unaffected and so the overall specification is improved by 20dB. The 7^{th} and 9^{th} order products have increased by 8dB and 10dB respectively.

Comparing figures 4.20 and 4.21 it can be seen that the 3^{rd} order products are cancelled to -53dBc an improvement of 34dB over the uncorrected case. The 5^{th} order products are unaffected and so the overall specification is improved by 20dB. The 7^{th} and 9^{th} order products have both increased by 6dB.

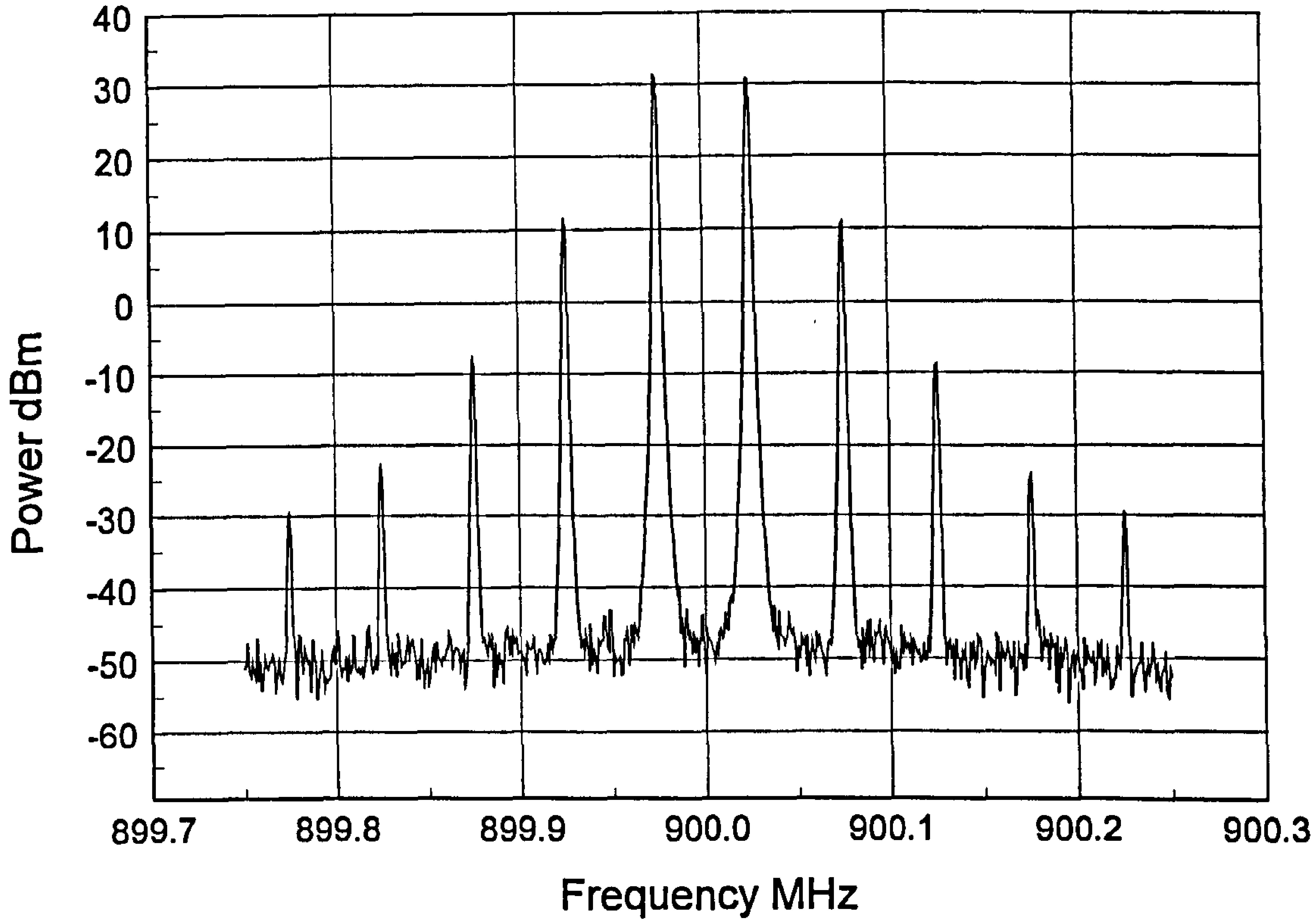

Figure 4.18 Plots of Two Tone Test at 50kHz Tone Spacing

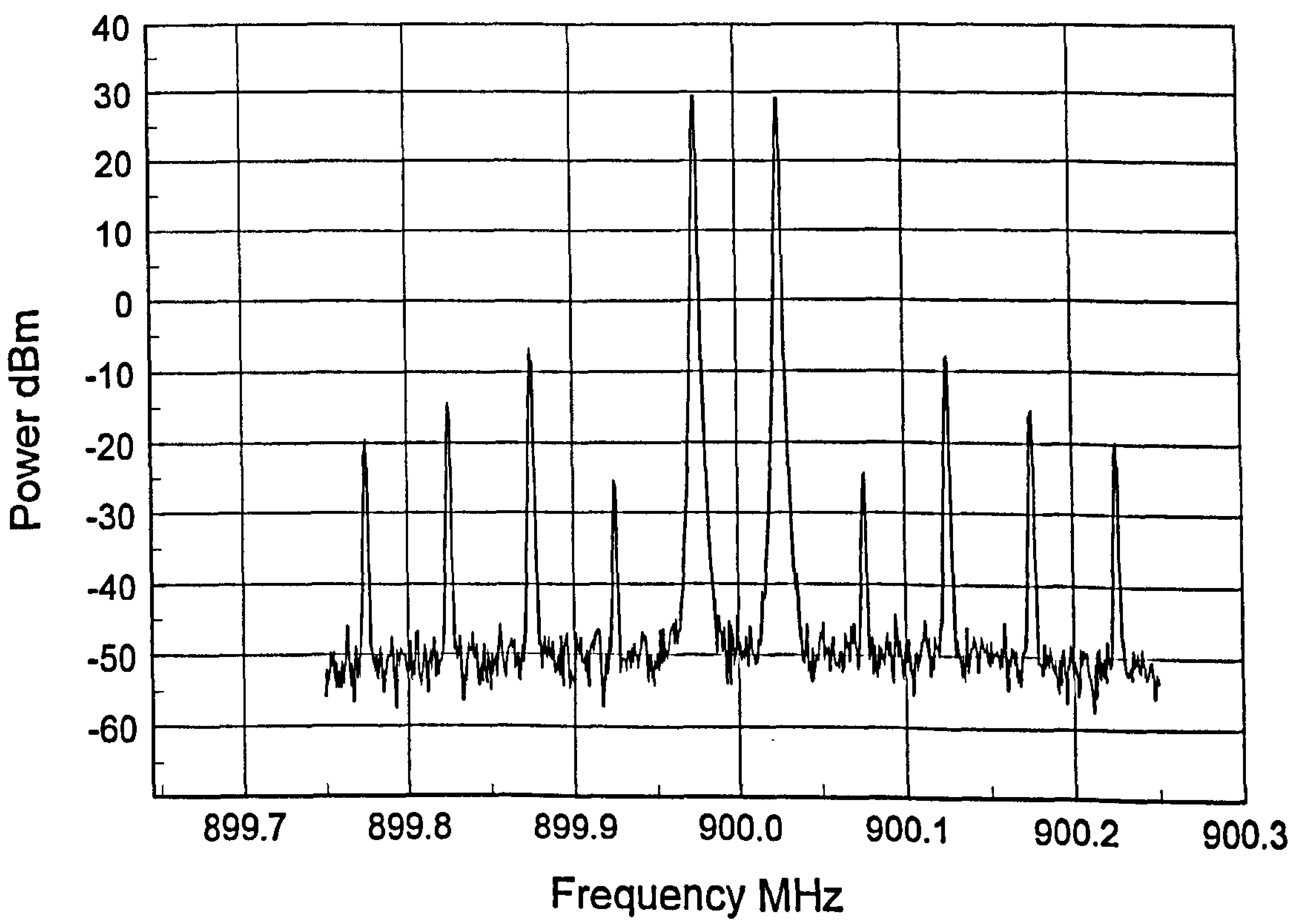

Figure 4.19 Plot of Predistorted Amplifier Output at 50kHz Tone Spacing

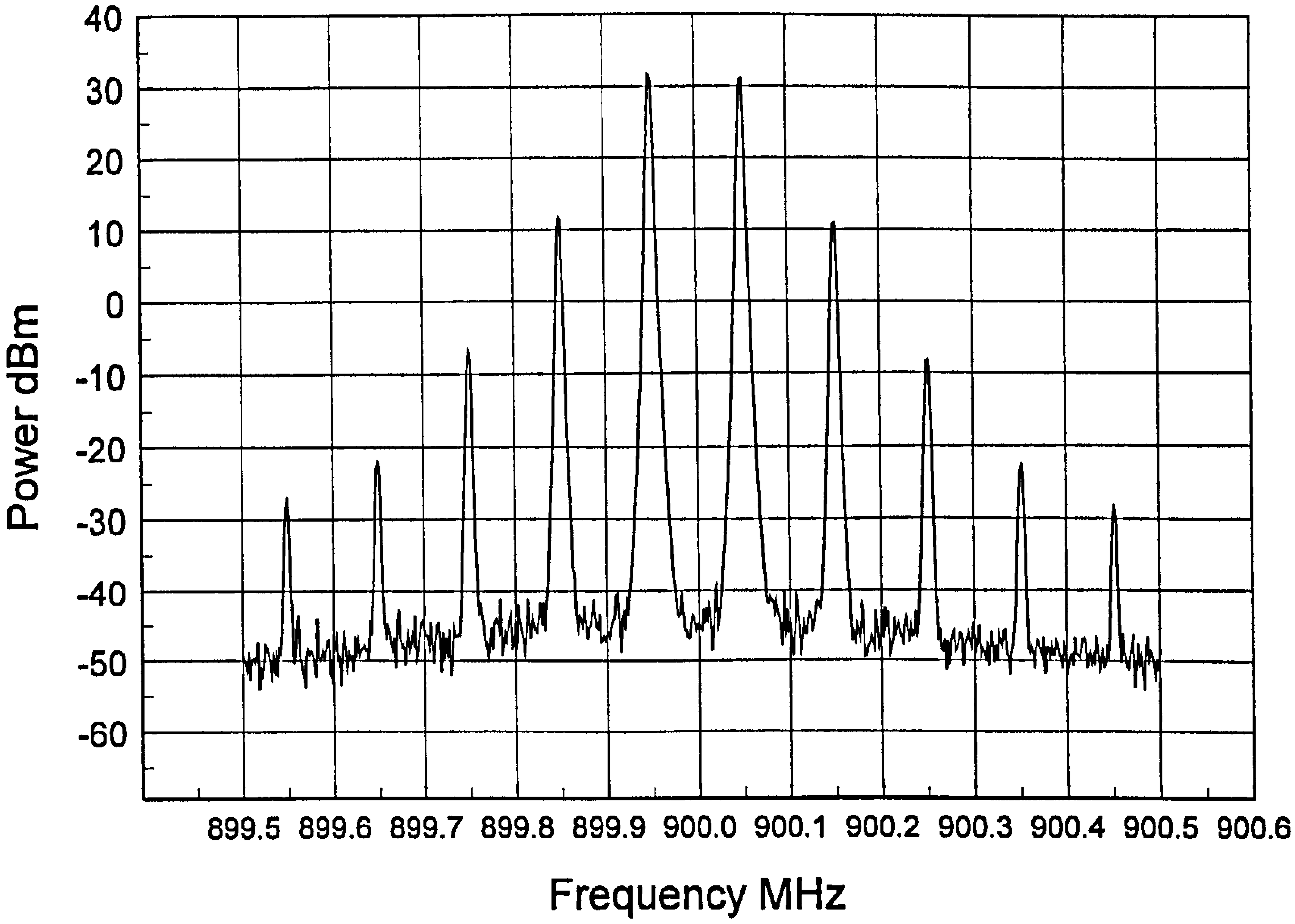

Figure 4.20 Plots of Two Tone Test at 100kHz tone Spacing

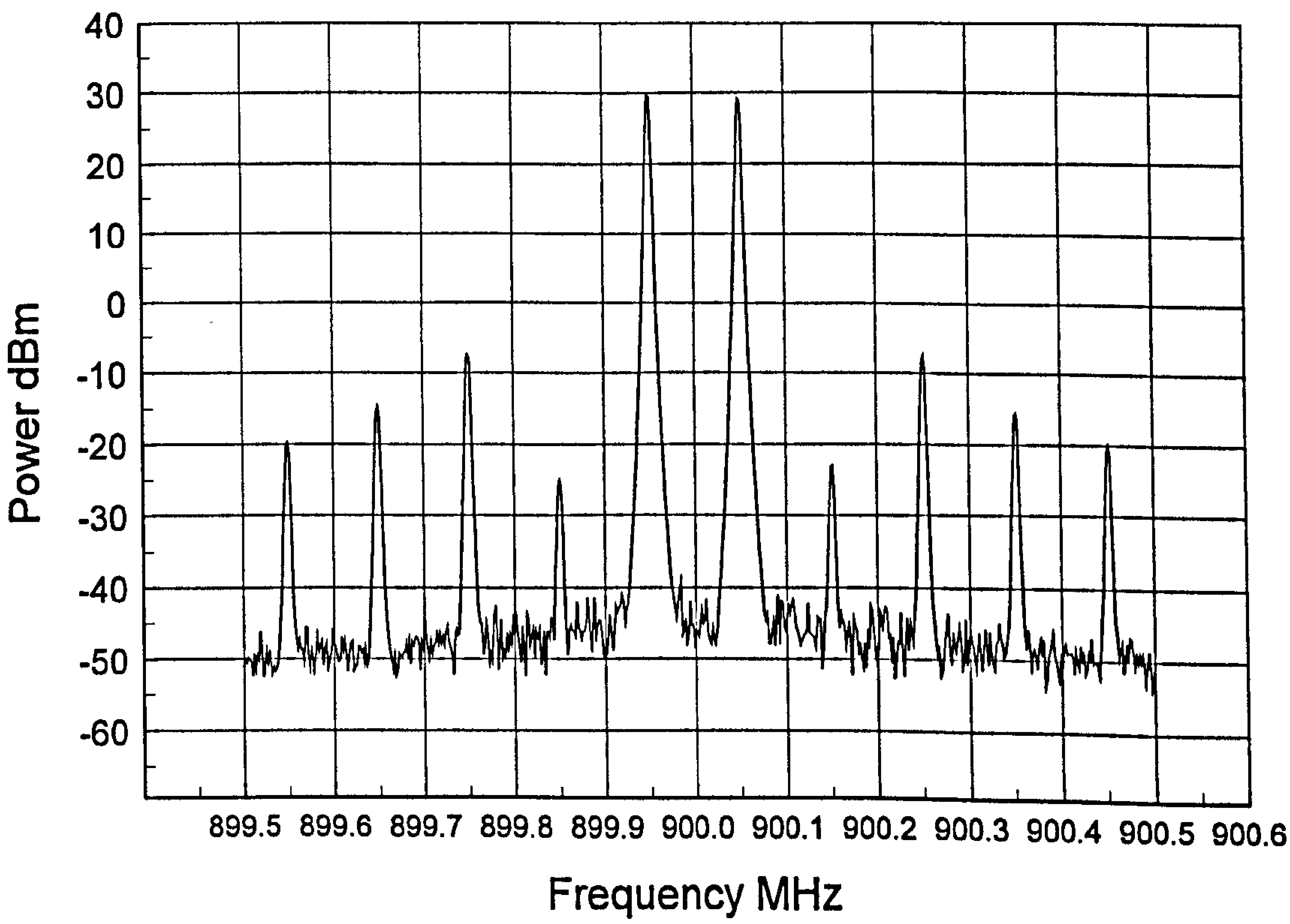

Figure 4.21 Plot of Predistorted Amplifier Output at 100kHz Tone Spacing

Comparing figures 4.22 and 4.23 it can be seen that the 3rd order products are cancelled to -38dBc an improvement of 18dB over the uncorrected case. The 5th order products however have increased slightly to -4dBm, this represents an increase of 1dB. This results in an overall improvement in specification of 14dB. The 9th order products have also increased by 10dB.

Comparing figures 4.24 and 4.25 it can be seen that the 3rd order products are cancelled to -35dBc an improvement of 15dB over the uncorrected case. The 5th order products however have increased, but not symmetrically the largest product having increased by 4dB. This results in an overall improvement in specification of 12dB. The 7th and 9th order products have also increased by 5dB and 13dB respectively.

The results [15] clearly show that this method of polynomial predistortion provides a good level of distortion reduction for narrowband, mediumband and broadband systems.

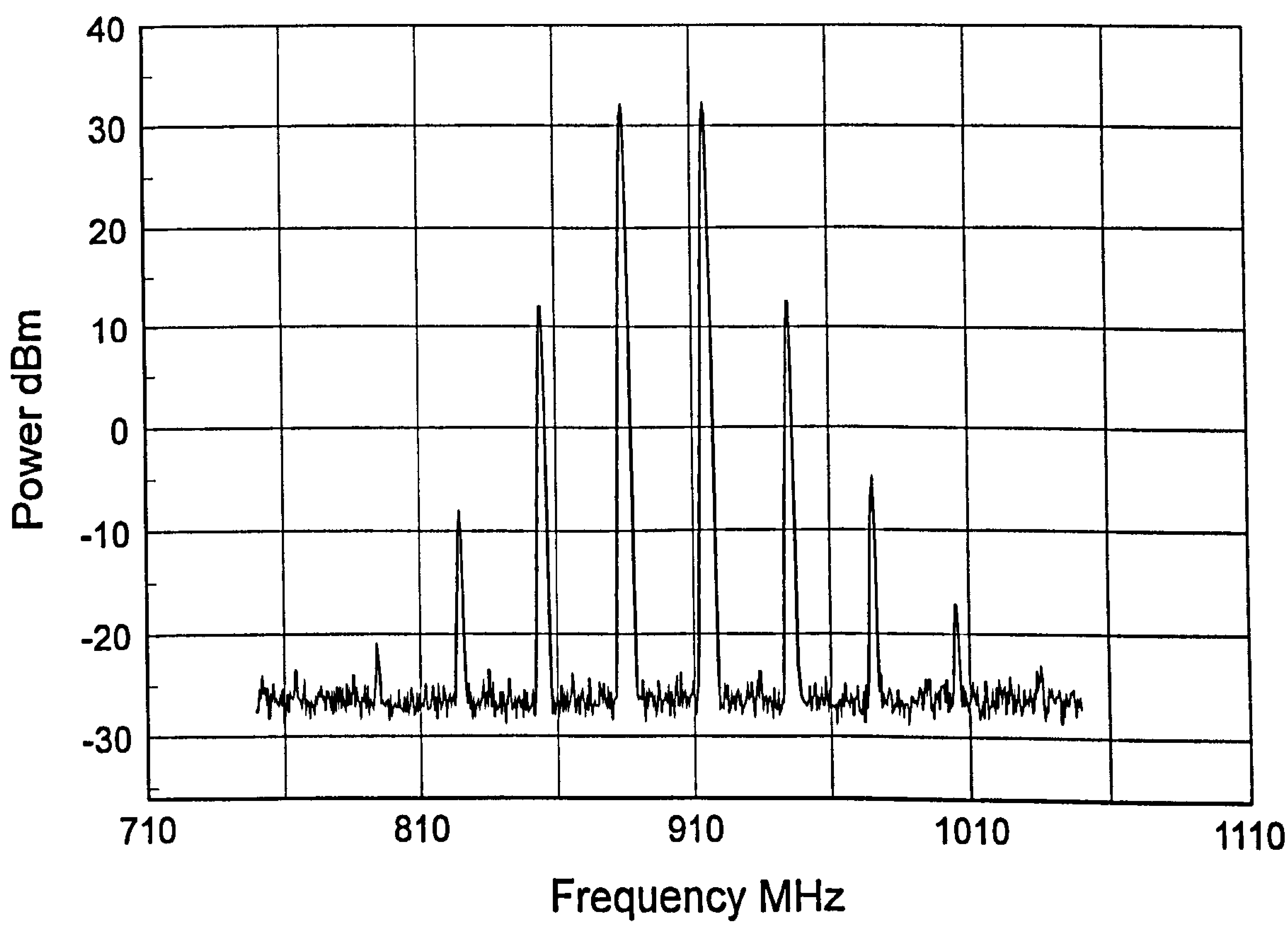

Figure 4.22 Plot of Two Tone Test at 30MHz Tone Spacing

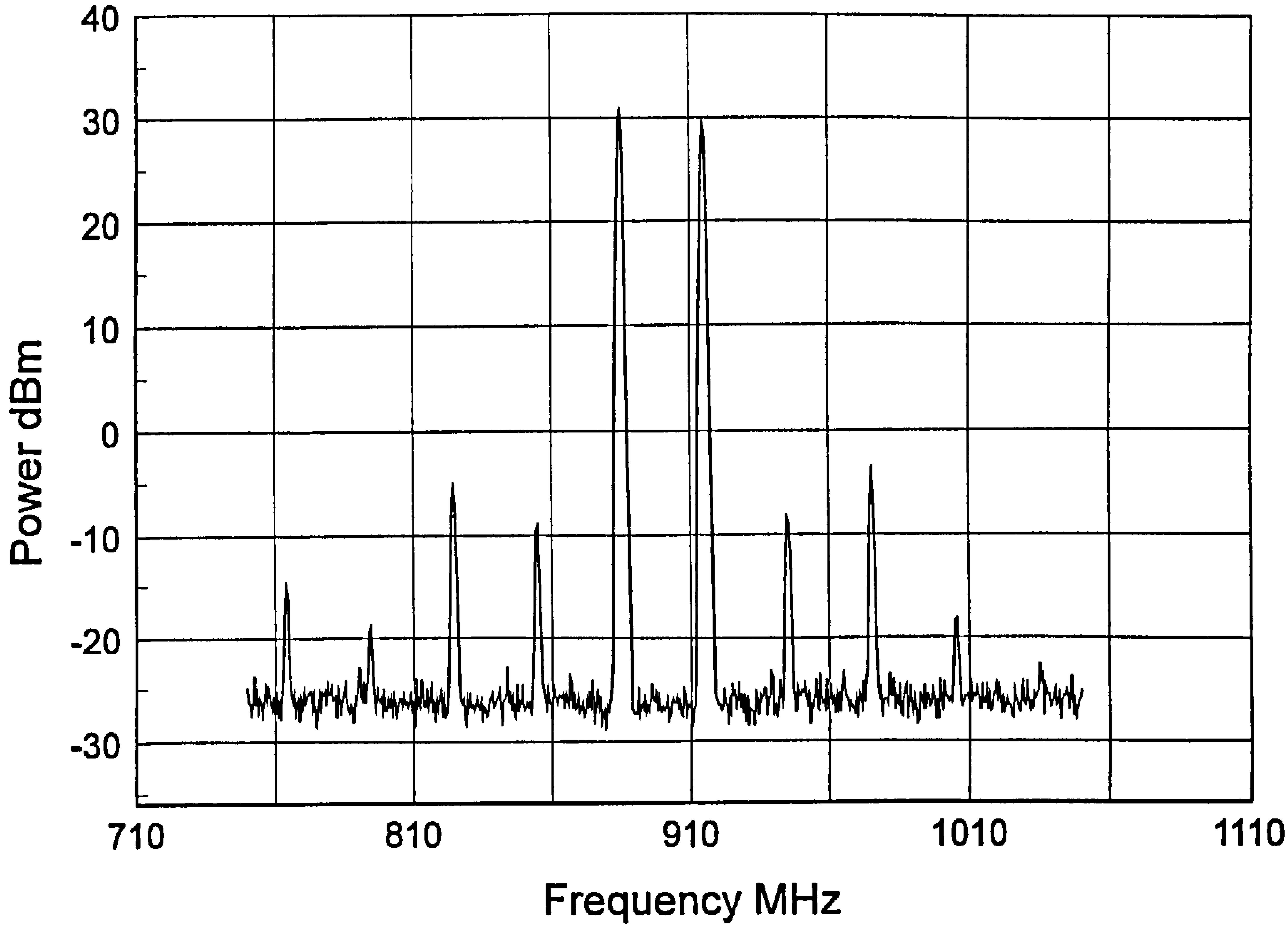

Figure 4.23 Plot of Predistorted Amplifier Output at 30MHz Tone Spacing

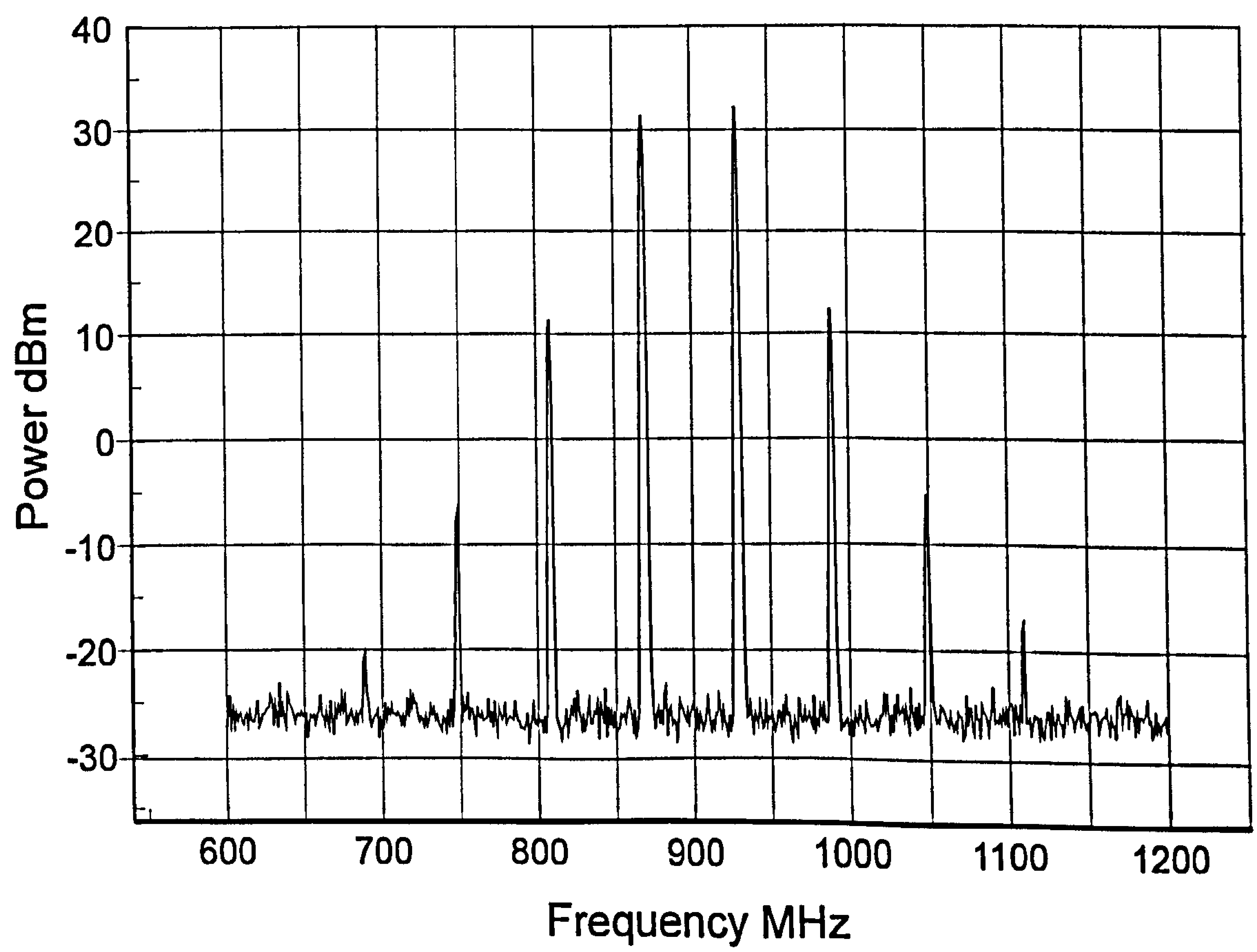

Figure 4.24 Plot of Two Tone Test at 60MHz Tone Spacing

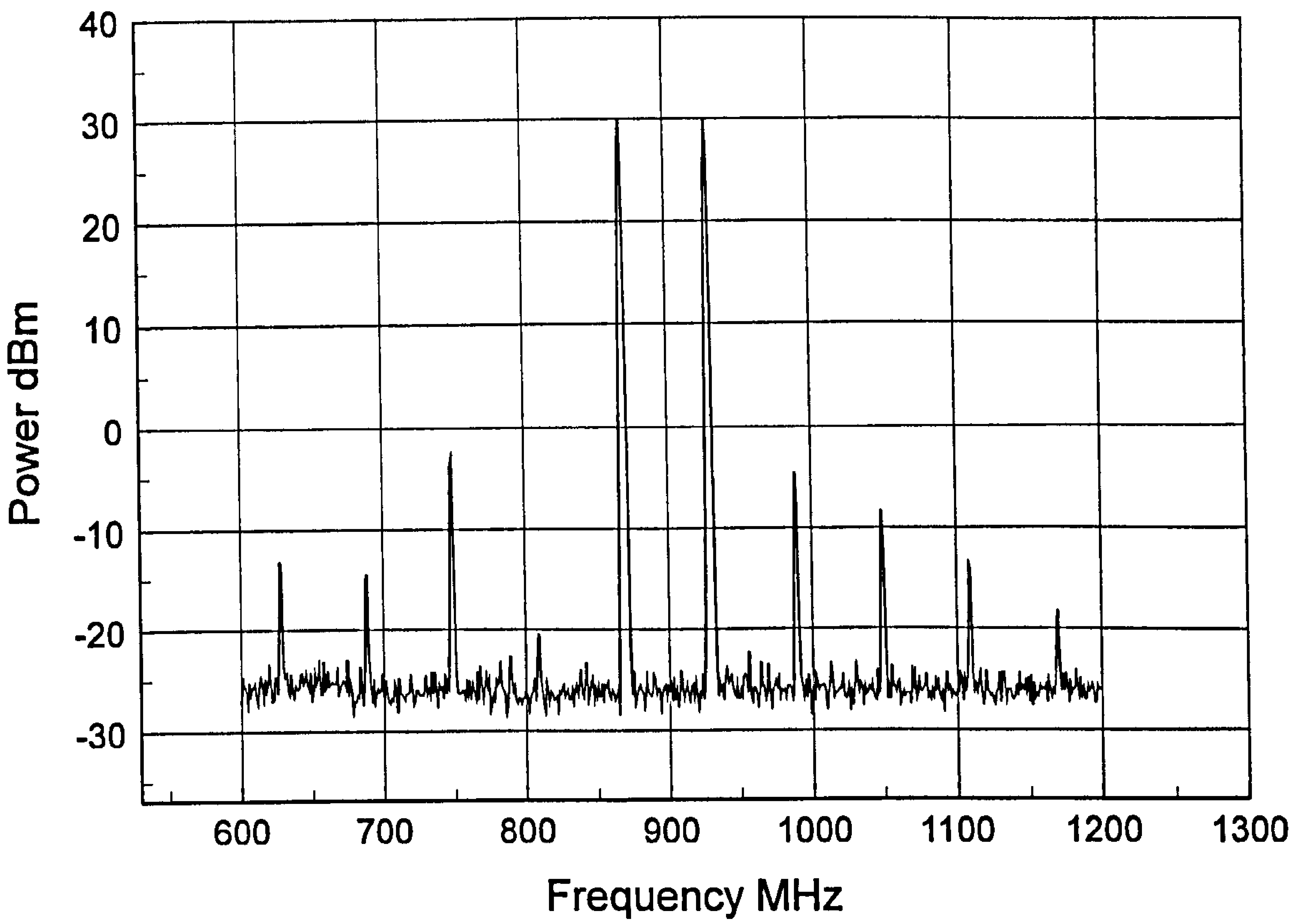

Figure 4.25 Plot of Predistorted Amplifier Output at 60MHz Tone Spacing

4.3.5 The Effects of Delay Error on Cancellation Performance

It has been shown thus far that improvements in the intermodulation performance of the amplifier used in this chapter are possible over very broad bandwidths indeed. It has been shown that good delay matching between the paths is important, but thus far no measurements of actual cancellation for particular delay errors have been made. The effect of delay on system performance has been investigated experimentally.

The effect of changes in predistorter delay were investigated as follows, the predistorter was set up for optimum cancellation at 900MHz with a tone spacing of 30MHz. The main path delay was then altered to introduce firstly too little delay and then too much delay. The delay measurements were then converted into their equivalent wavelength errors, the final result is shown in figure 4.26.

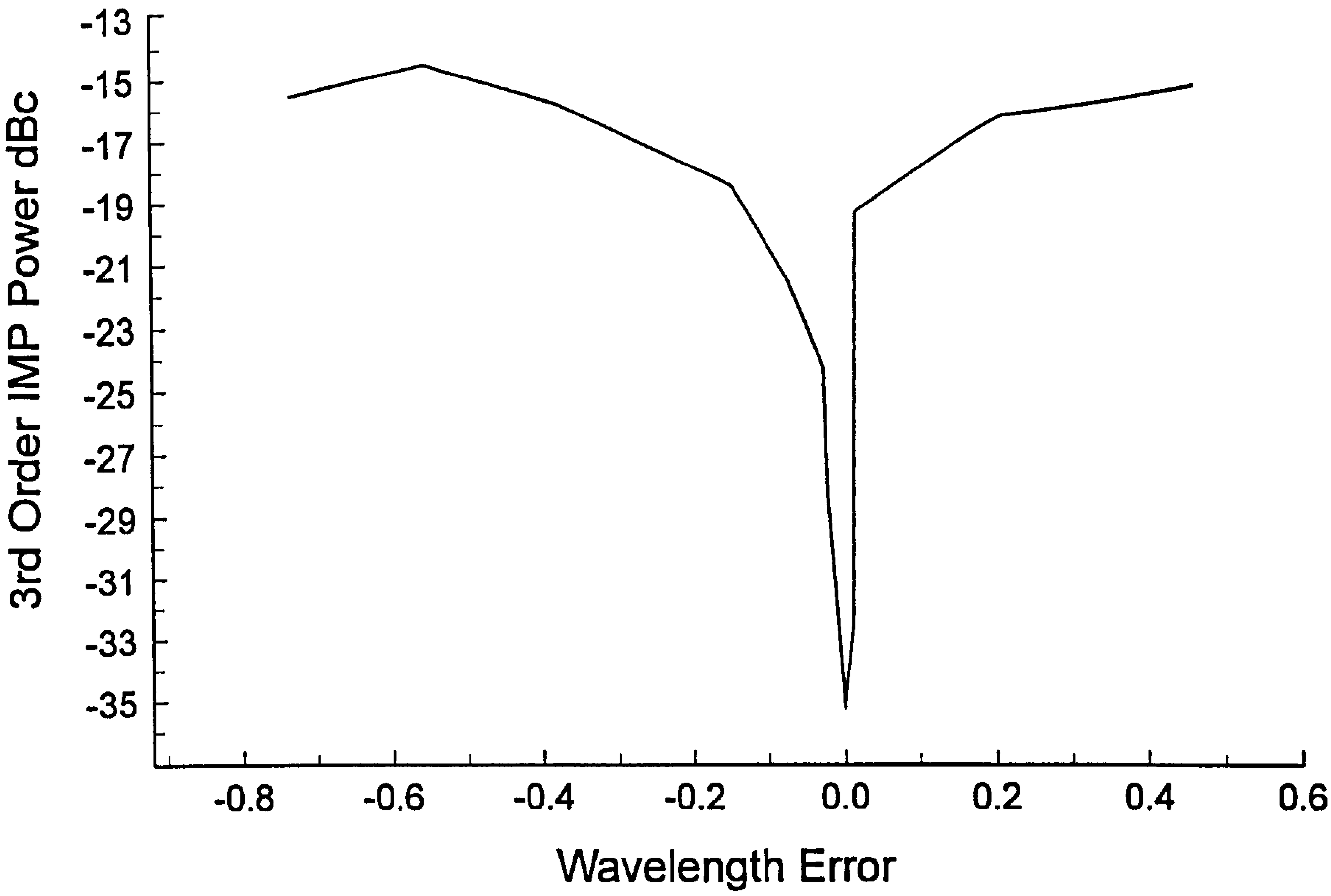

Figure 4.26 Graph of Wavelength Error vs 3rd IMD Power

Figure 4.26 shows how critical correct matching of the main and predistorted delays are to the cancellation performance of the predistorter. The graph shows clearly how the cancellation performance drops off rapidly even for small amounts of delay error i.e. errors greater than +/- 0.05λ cause significant reductions in cancellation performance.

4.3.6 Practical Measurements of Gain and Phase Error

In order to verify the theoretical analysis carried out in chapter 3 measurements were taken for various gain and phase errors. The experimental set up is shown in figure 4.27. The predistortion system consisted of a cubic non-linearity, variable gain and phase adjustment, a main path attenuator, an input splitter and an output combiner. The amplifier chosen for the tests was a MAV 11 which is a MIMIC amplifier with a predominantly third order characteristic [16] supplied by mini circuits. This amplifier was chosen because it most closely mimics the third order only type of non-linearity used in the analysis of chapter 3.

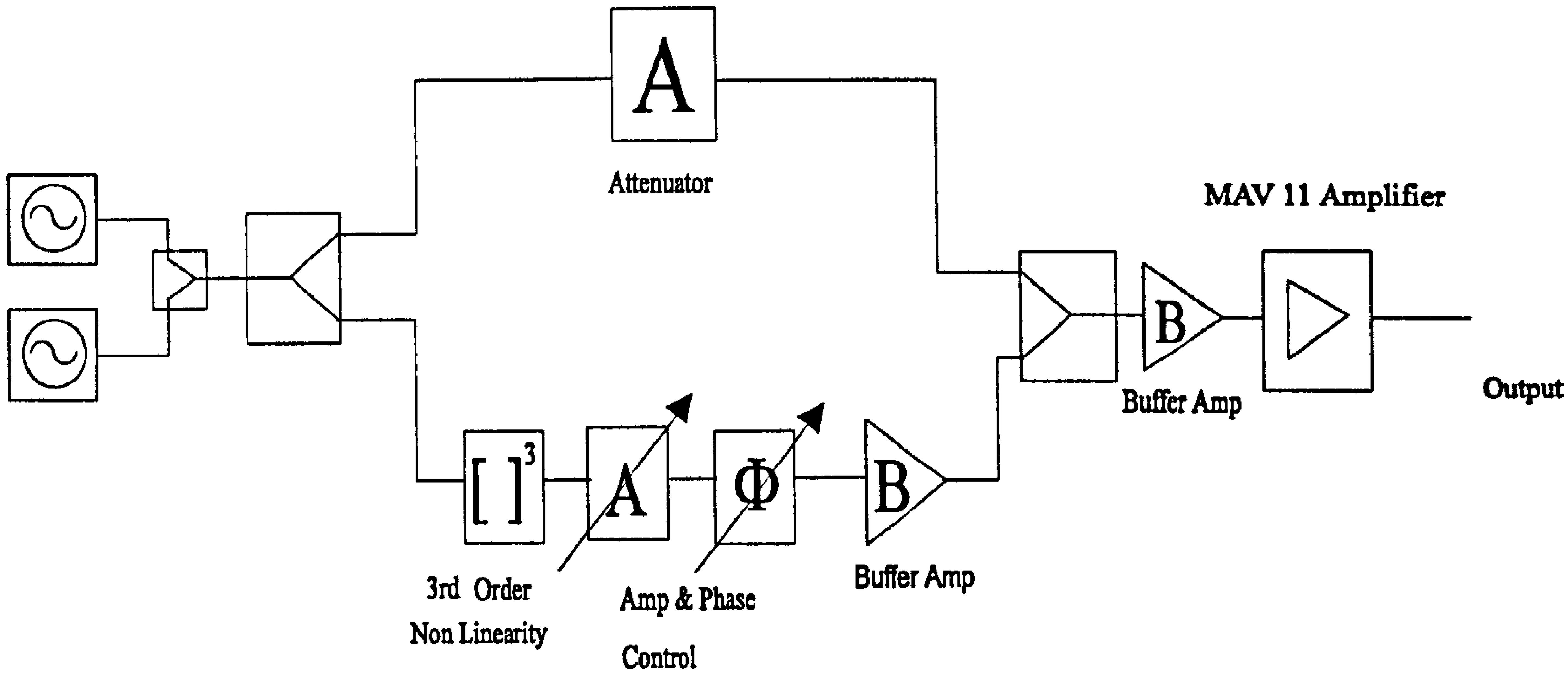

Figure 4.27 Experimental set up for Gain and Phase Error Measurements

The following procedure was followed when taking the measurements, firstly the system was set up for optimal cancellation while driving the amplifier at its 1dB compression point. Then measurements were taken of cancellation while a range of constant attenuation errors was applied while the phase error was varied. Then a range of constant phase errors was applied while the attenuation error was varied. The actual gain and / or phase error applied at each step was measured using a HP8753E network analyser to ensure that neither the attenuator introduced additional phase error or the phase shifter introduced additional attenuation error. The results of these measurements are shown in figures 4.28 and 4.29 respectively.

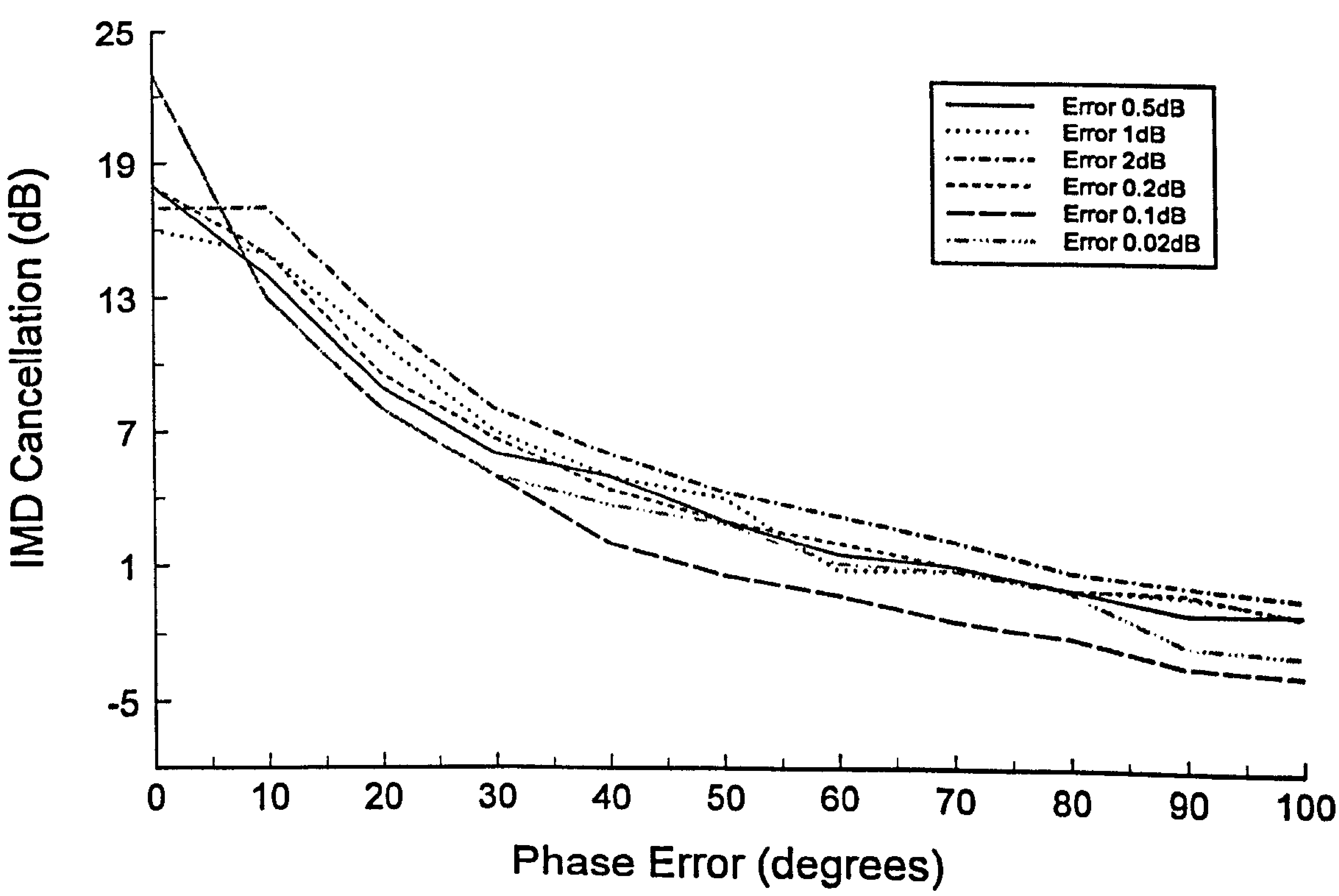

Figure 4.28 Measurements of IMD Cancellation for Various Gain Errors

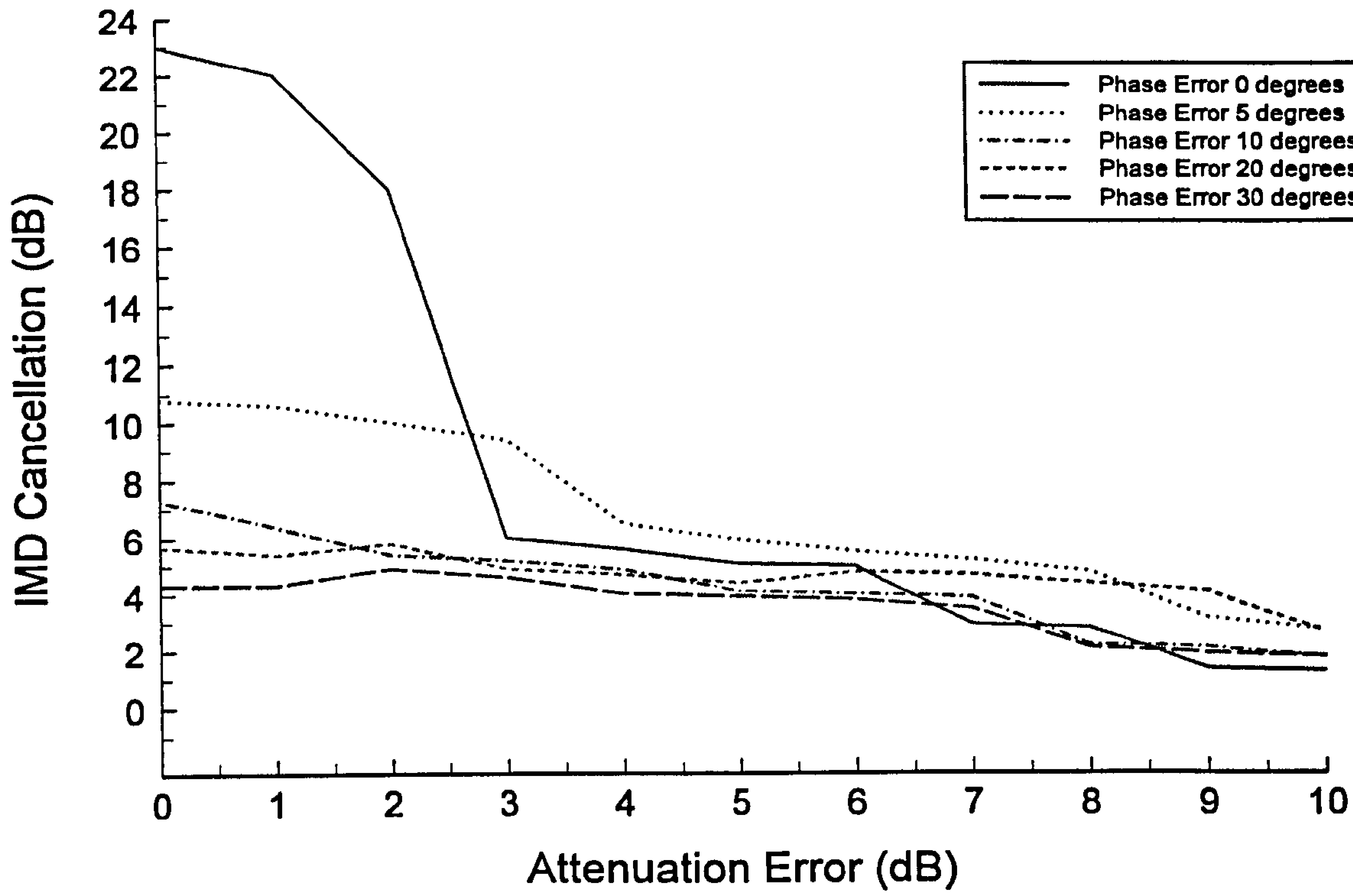

Figure 4.29Measurements of IMD Cancellation for Various Phase Errors

The results in figures 4.28 and 4.29 show quite clearly how the system performance degrades quite significantly with slight changes in attenuation error. The overall system appears to be slightly more tolerant of phase errors. With reference to figure 4.28 it may be observed that the resultant IMD cancellation is very dependent on the gain error when the gain error is 0.02dB, for all other gain errors the main factor in performance is the phase error. Once the phase error exceeds 10°then the cancellation performance is very dependent on the phase error. Referring to figure 4.29, for the case when the phase error is 0° it can be seen that the performance degrades rapidly as gain error increases from 0dB to 3dB. When the phase is increased to 5°a very substantial fall in performance results, at 0dB gain error the cancellation performance is reduced by 12dB when compared to the case for 0°phase error. Once the phase error is greater than 5°then performance levels out with an average cancellation of 5dB no matter what the gain error.

4.3.7 Control Issues Associated with Cubic Polynomial Predistorters

Automatic control of this predistortion system is possible due to the use of voltage-controlled attenuators and phase shifters in the predistorted path. The system has been investigated to see if the added complexity of a control scheme is justified. The level of the IMP's relative to the main tones was measured for various levels of input power. The measurements were undertaken at a centre frequency of 900MHz using a predistortion system optimised for operation at wide bandwidths, with a tone spacing of 30MHz being chosen for all measurements. These measurements were undertaken for three sets of conditions firstly the amplifier's IMD performance was measured with no correction system connected. Then measurements of intermodulation performance were undertaken for various levels of input power while the predistortion system was manually adjusted for the lowest level of 3rd order IMP possible. Finally the system was set-up for optimum cancellation at the amplifiers 1dB compression point, the amplifier input power was then backed off without any further adjustment of the attenuators and phase shifters while the IMP performance was monitored. The results of these measurements are shown in figure 4.30.

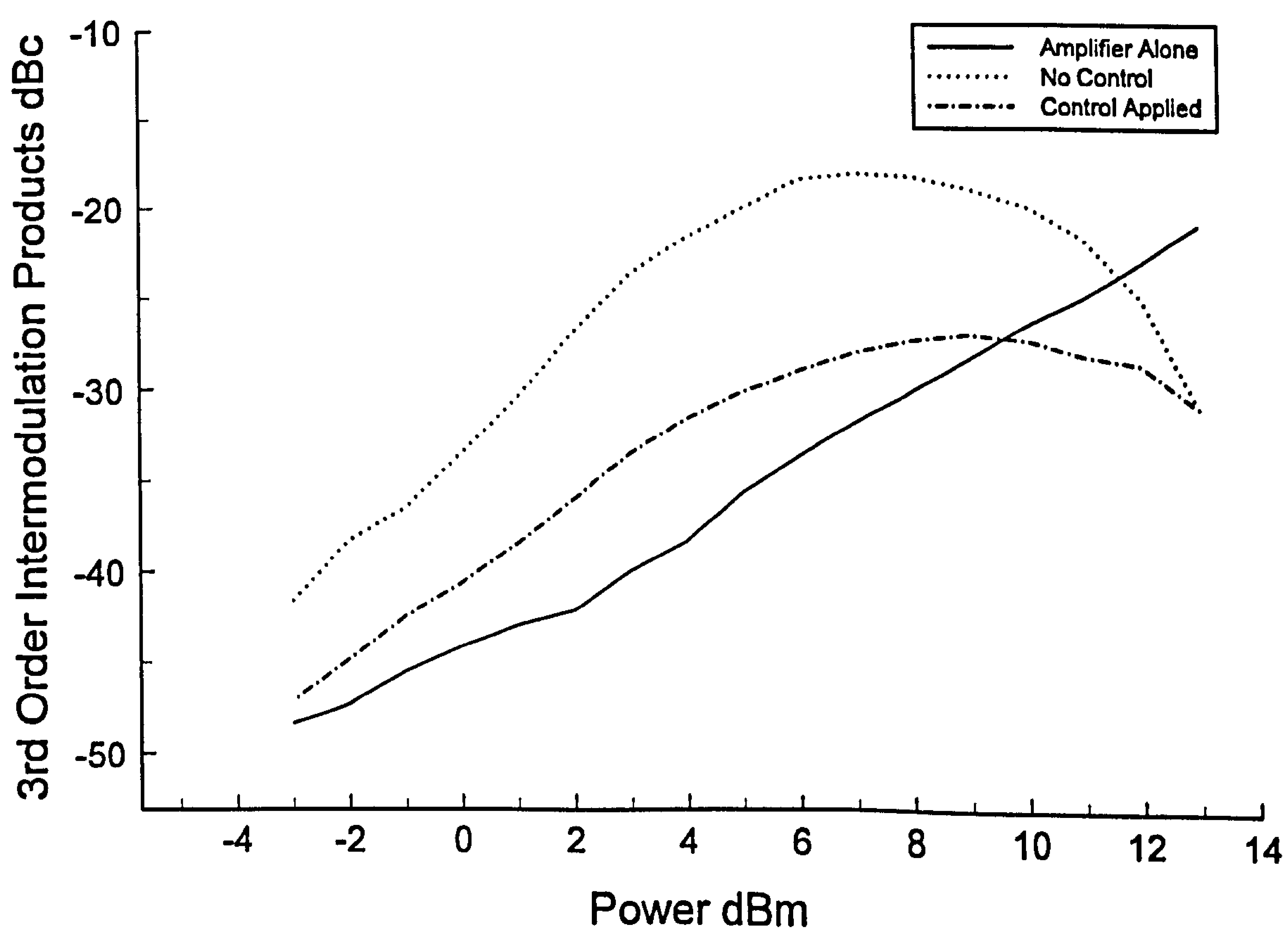

Figure 4.30 Measurements of Amplifier and Predistorted Amplifier System Back-off

It may be seen from figure 4.30 that if the amplifier is backed off without any linearisation scheme applied the amplifier intermodulation performance gradually improves as the output power to the amplifier is reduced. This is as expected for an amplifier of this type because as back-off increases then the more the amplifier is being operated within it's linear region of operation i.e. the level of the intermodulation products will fall away. Now considering the plots where the predistorter is present now the situation is less clear. If the predistortion system is adjusted for optimum operation point then initially the predistorter performs much better than the amplifier alone. But if the amplifier input power is 9dBm or less then the amplifier alone performs better than the predistortion system. The reason this occurs is simply due to the fact that the amplifier is now predominantly operating in it's linear region and so the polynomial coefficient being added into the amplifier now is effectively reducing the amplifiers performance and not enhancing it. Now if we finally consider the case where we set up the predistorted amplifier system to operate for minimum IMD and then back off the input power without altering the attenuation or phase adjustment applied, it can be seen that the performance degrades more rapidly and the performance is actually much worse than the amplifier alone once the input power is less than 11dBm which is equal to 2dB of back off.

It would appear form these results that it is worthwhile adding some form of control scheme to control the system for various levels of input power. However the mixer based predistortion system requires an input power greater than 2dBm in order to keep the conversion loss of the mixers low enough for the predistorter to generate a good quality cubic output. If the input to the predistorted amplifier system drops below 7dBm then the input power to the cubic element will fall to a level that is below the point for good cubic element operation. This results in the situation shown in figure 4.30 where the predistorter actually degrades the amplifier performance rather than enhancing it. Secondly an amplifier which has a predominantly 3rd order characteristic as in this case will generally operate as a class A or a class AB amplifier. These classes of amplifier exhibit their best efficiency performance when they are driven at their 1dB compression point. So driving the amplifier at a lower level of output power by backing off the amplifier is undesirable since it will have a negative effect on the efficiency of the amplifier being linearised.

4.3.7.1 Transfer Characteristics of the Cubic Polynomial Predistortion System

As has been stated in chapter 3 the aim of any predistortion system is to generate a transfer characteristic that is the inverse of the amplifier transfer characteristic. If measurements are made of the amplifier, the predistorter and the whole-predistorted system amplitude and phase responses then it may be shown how well or badly the predistorter generates an inverse to the amplifiers transfer function. The amplitude and phase information, which is generated in this way, may be converted into the in-phase (I) and the quadrature (Q) performance of the system. Measurements were taken of the amplifier, predistorter and predistorter plus amplifier gain and phase response these are shown in figures 4.31, 4.32 and 4.33 respectively. Measurements were also taken of the amplifier, predistorter and predistorter plus amplifier voltage transfer functions and the I & Q transfer functions the results are shown in figures 4.34, 4.35, 4.36, 4.37, 4.38 and 4.39 respectively.

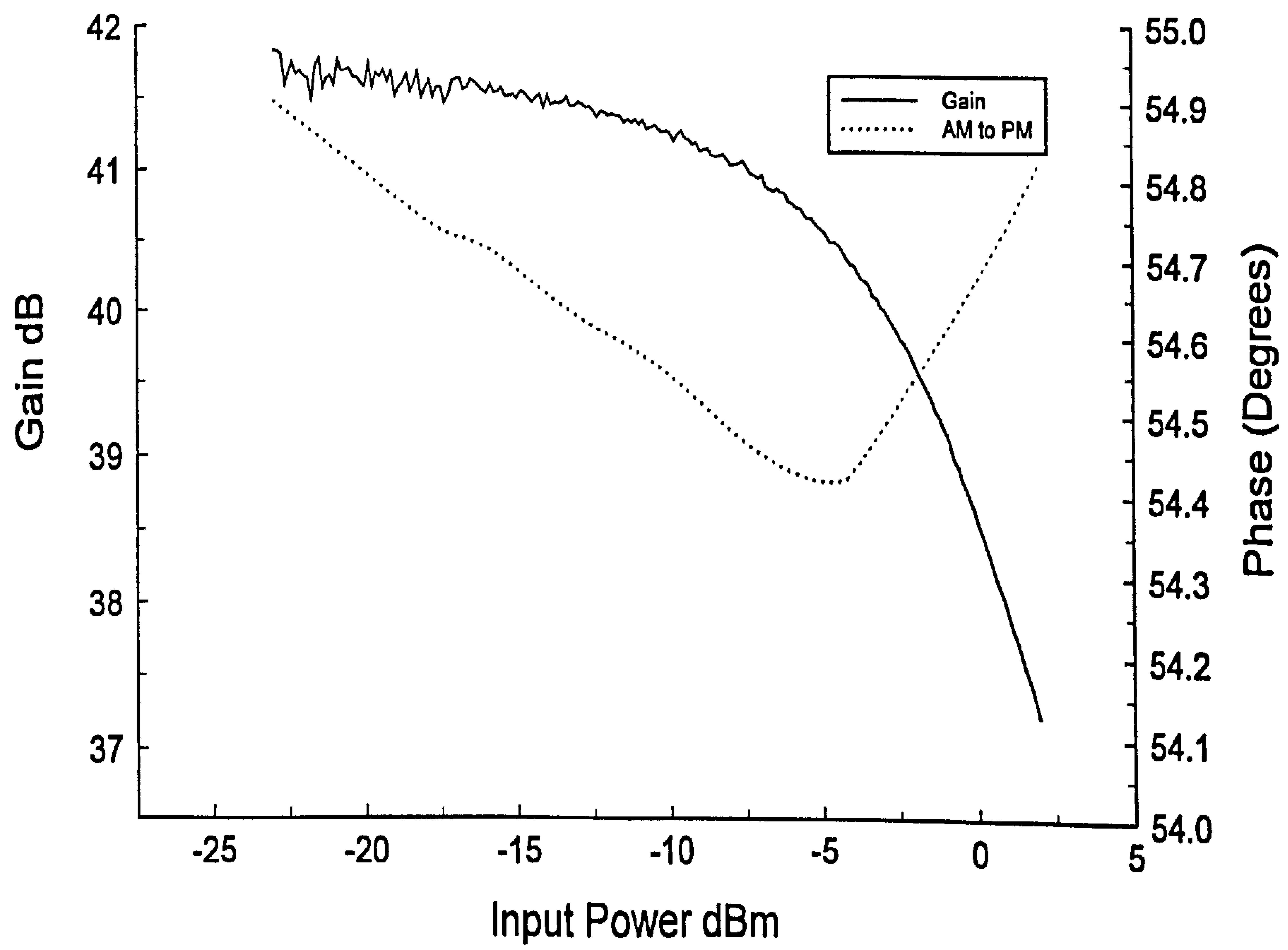

Figure 4.31 Amplifier Gain and Phase Response

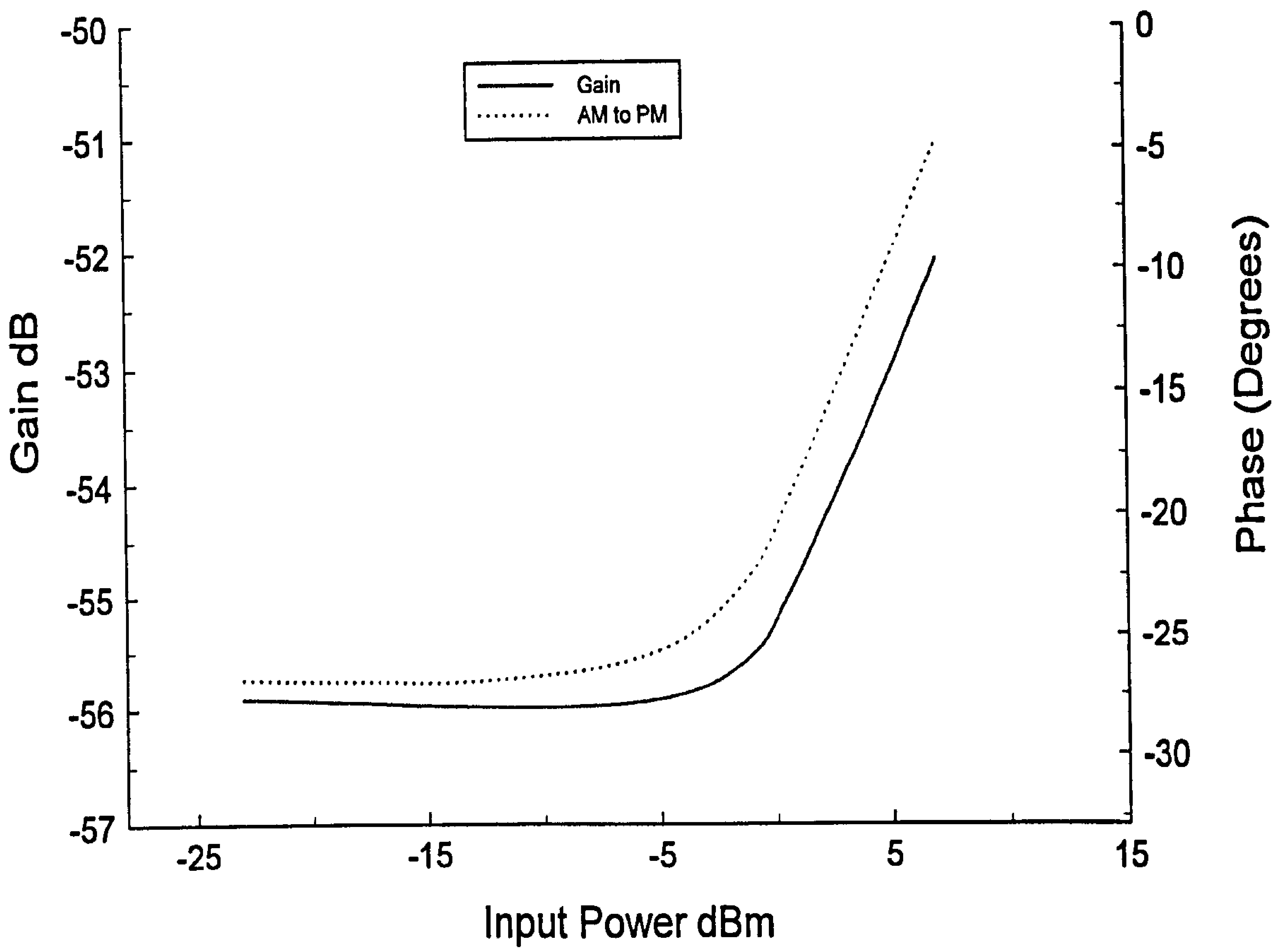

Figure 4.32 Cubic Predistorter Gain and Phase Response

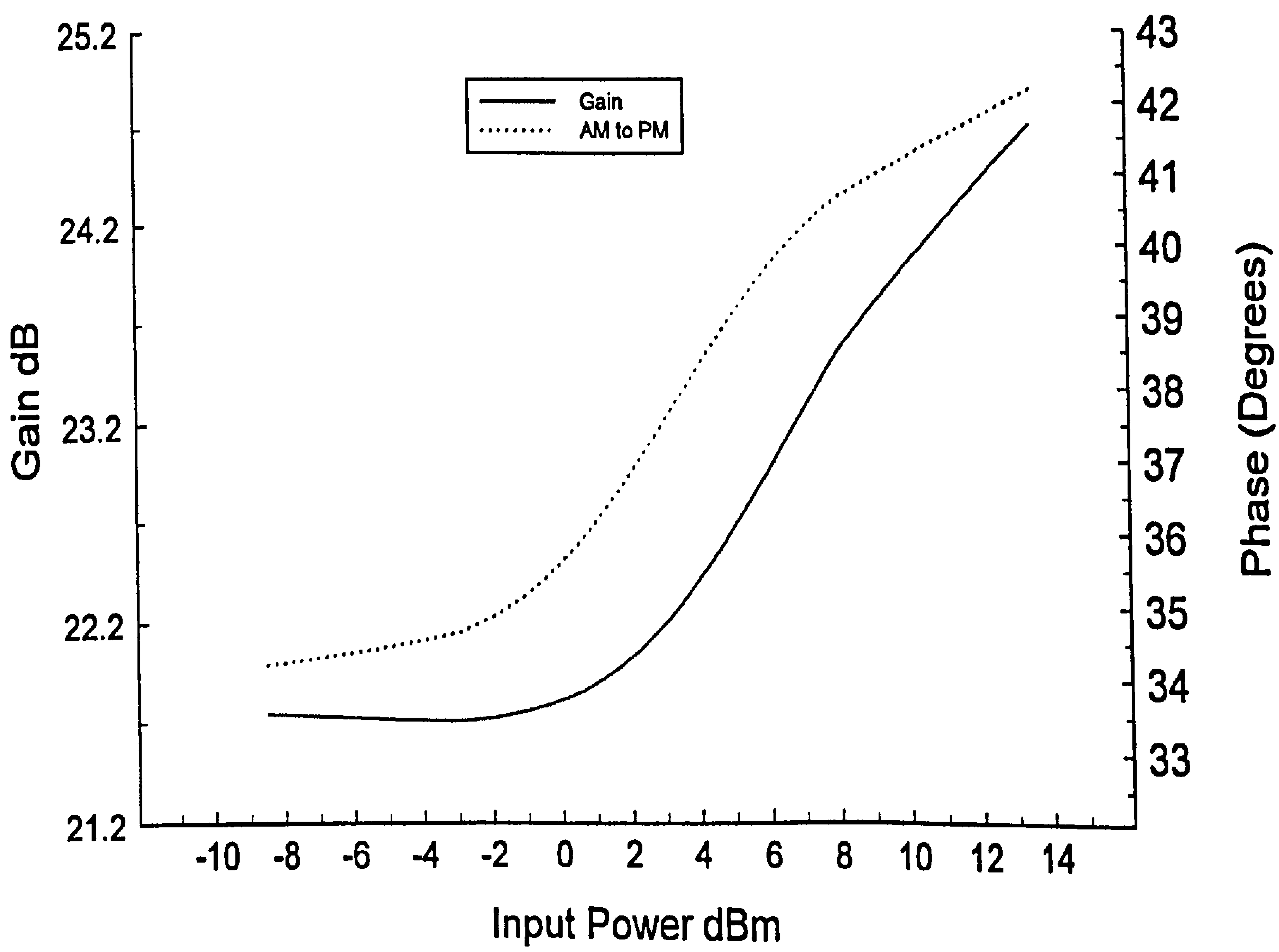

Figure 4.33 Cubic Predistortor plus Amplifier Gain and Phase Response

Figure 4.32 shows how the predistorter gain at very low power levels is again flat and shows how as the power level is increased the gain becomes very expansive. Figure 4.31 shows that for the amplifier the opposite is true with a short gain flat region followed by a long compressive region. The resultant effect is an extension to the region of gain flatness followed by a region of gain expansion finally followed by a region of gain compression these effects are shown in figure 4.33. The final region of gain compression is due to the system being driven to the point of excessive gain compression, into the area where the amplifier would not normally be operated.

Comparing figures 4.31 and 4.32 the phase response shows some inverse characteristic but the amplifier only introduces very limited phase change over all input powers. The predistorter however introduces very little phase change at low power but a fairly large phase change as the power levels are increased. This results in sub-optimal overall phase response that may be seen in figure 4.33.

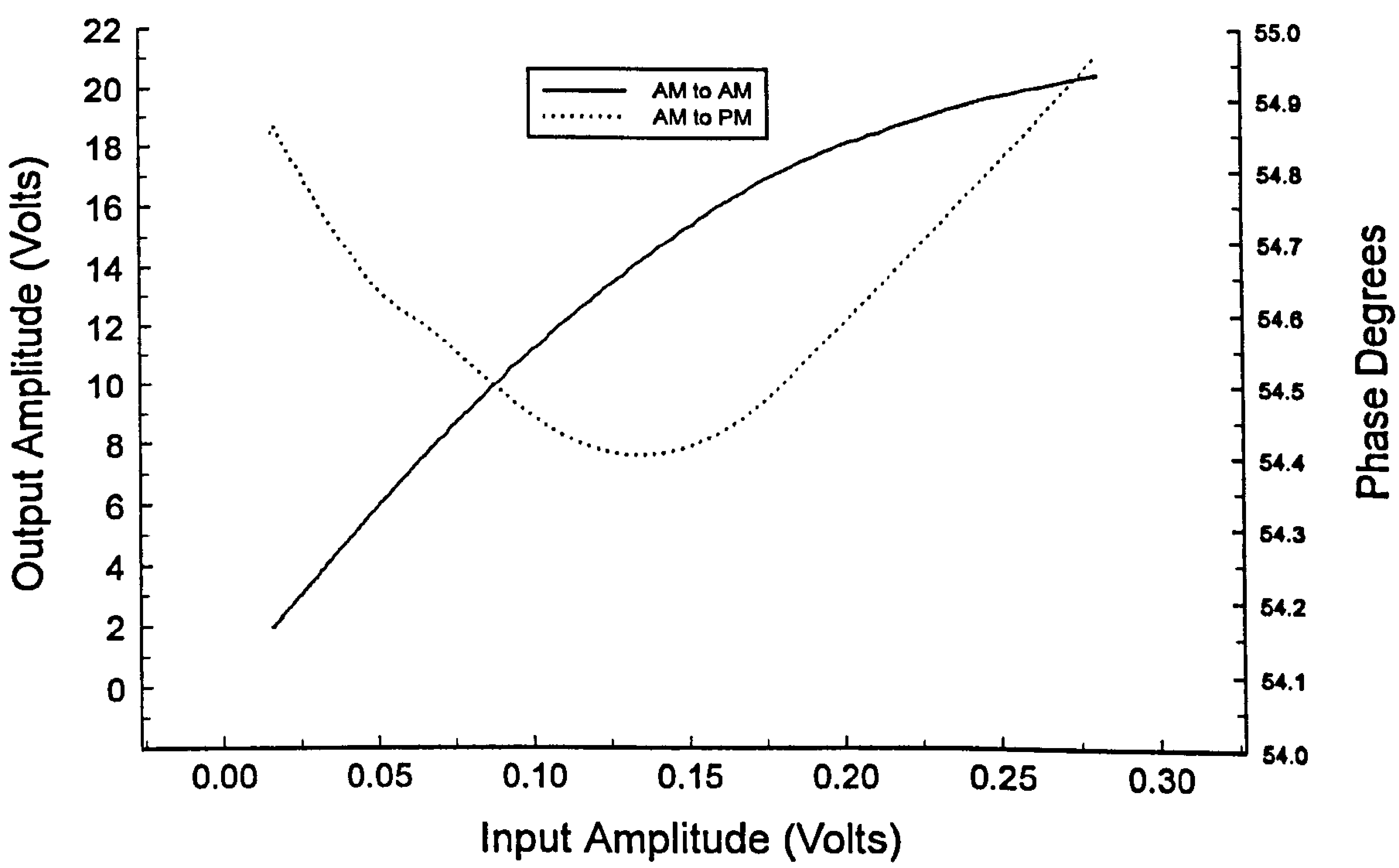

Figure 4.34 Amplifier Voltage Transfer Function

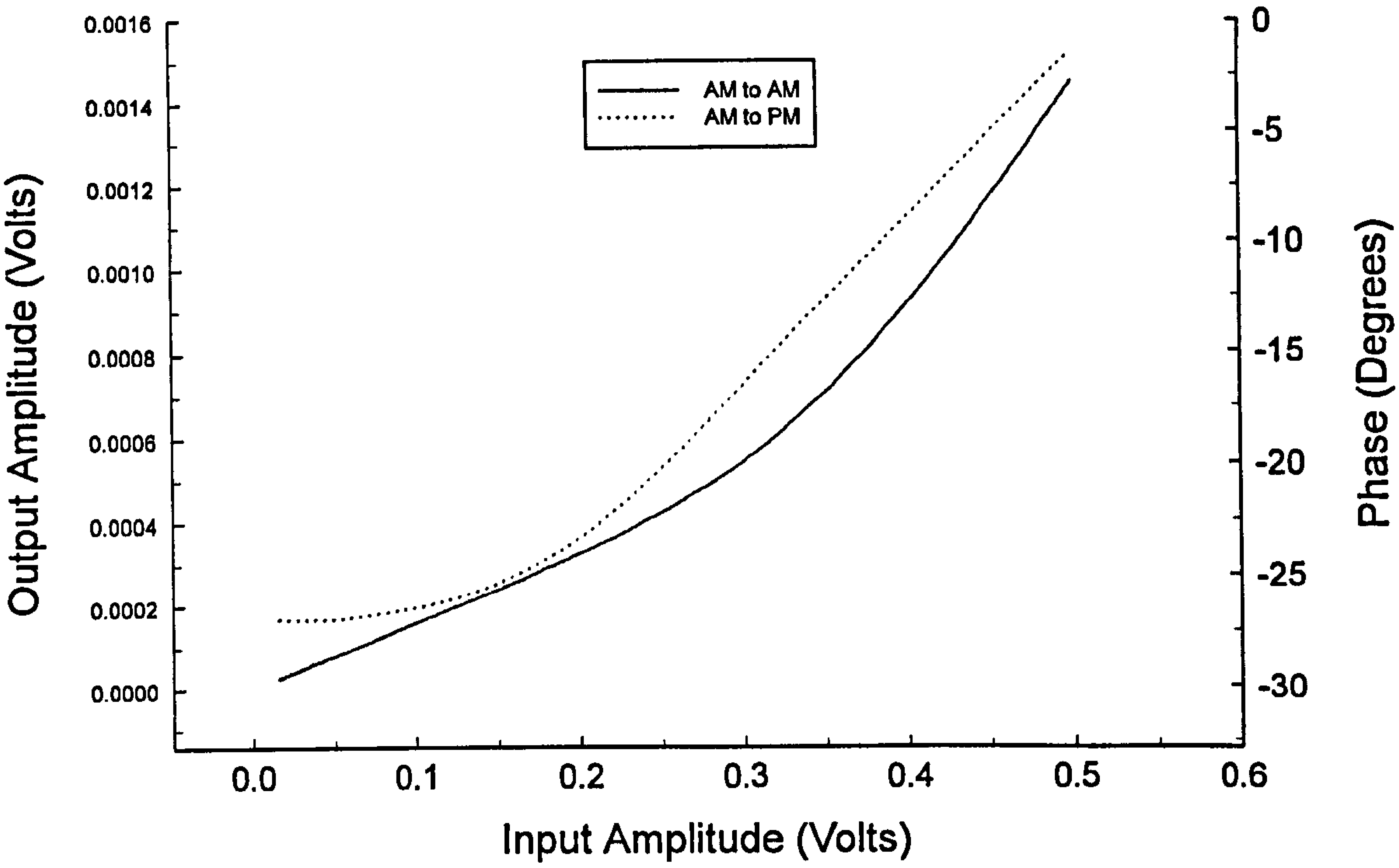

Figure 4.35 Cubic Predistortor Voltage Transfer Function

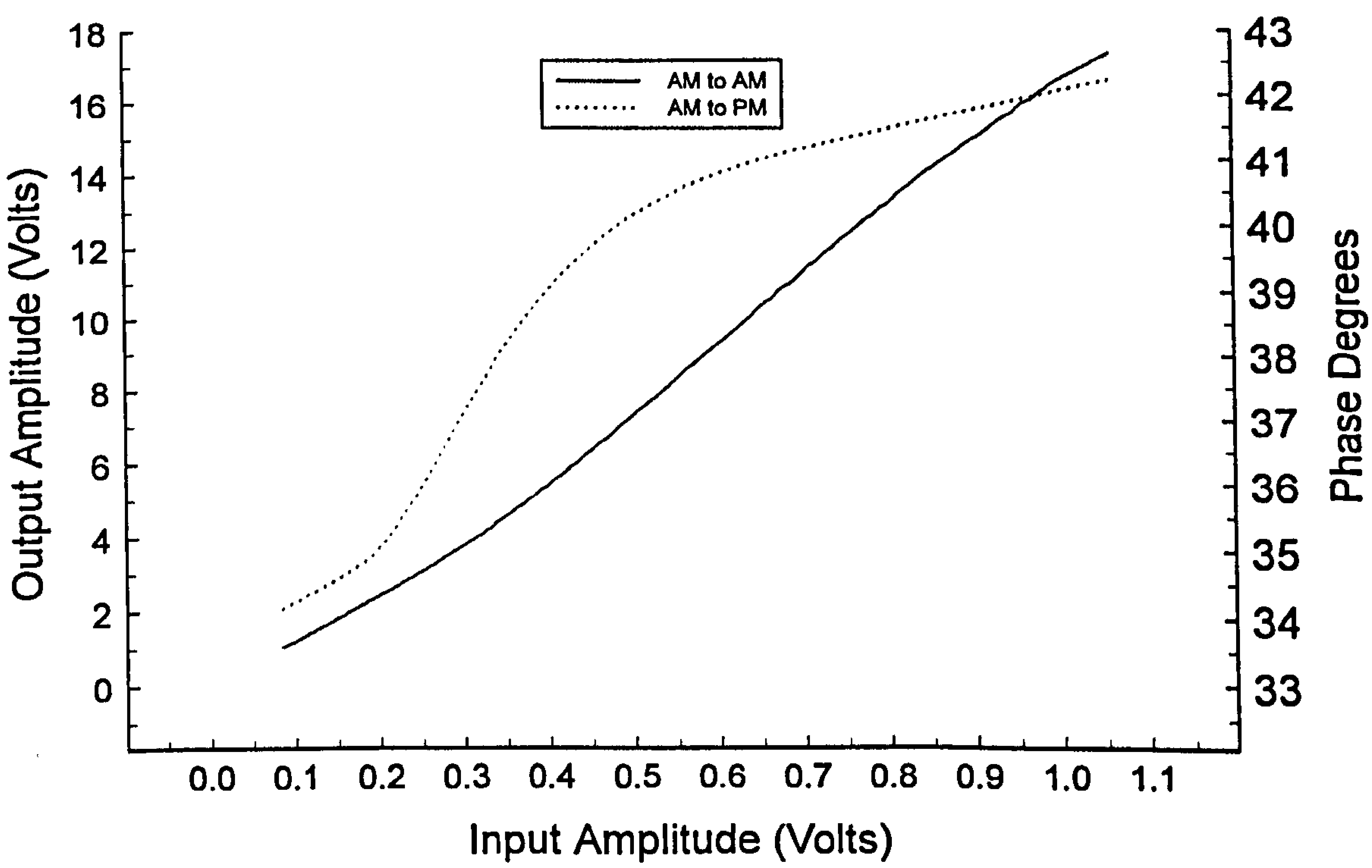

Figure 4.36 Cubic Predisorter plus Amplifier Voltage Transfer Function

The voltage transfer function of the amplifier shown in figure 4.34 shows how the amplifier has a compressive AM to AM characteristic and a relatively gentle AM to PM characteristic. The cubic predistorter has an expansive AM to AM characteristic and a more severe AM to PM characteristic than the amplifier this can be seen in figure 4.35. The resultant combination of the amplifier and the predistorter is shown on figure 4.36. The AM to AM characteristic can be seen to be much more linear than the amplifier was on it's own. The effect of the excessive amount of AM to PM correction introduced by the predistorter can also be seen.

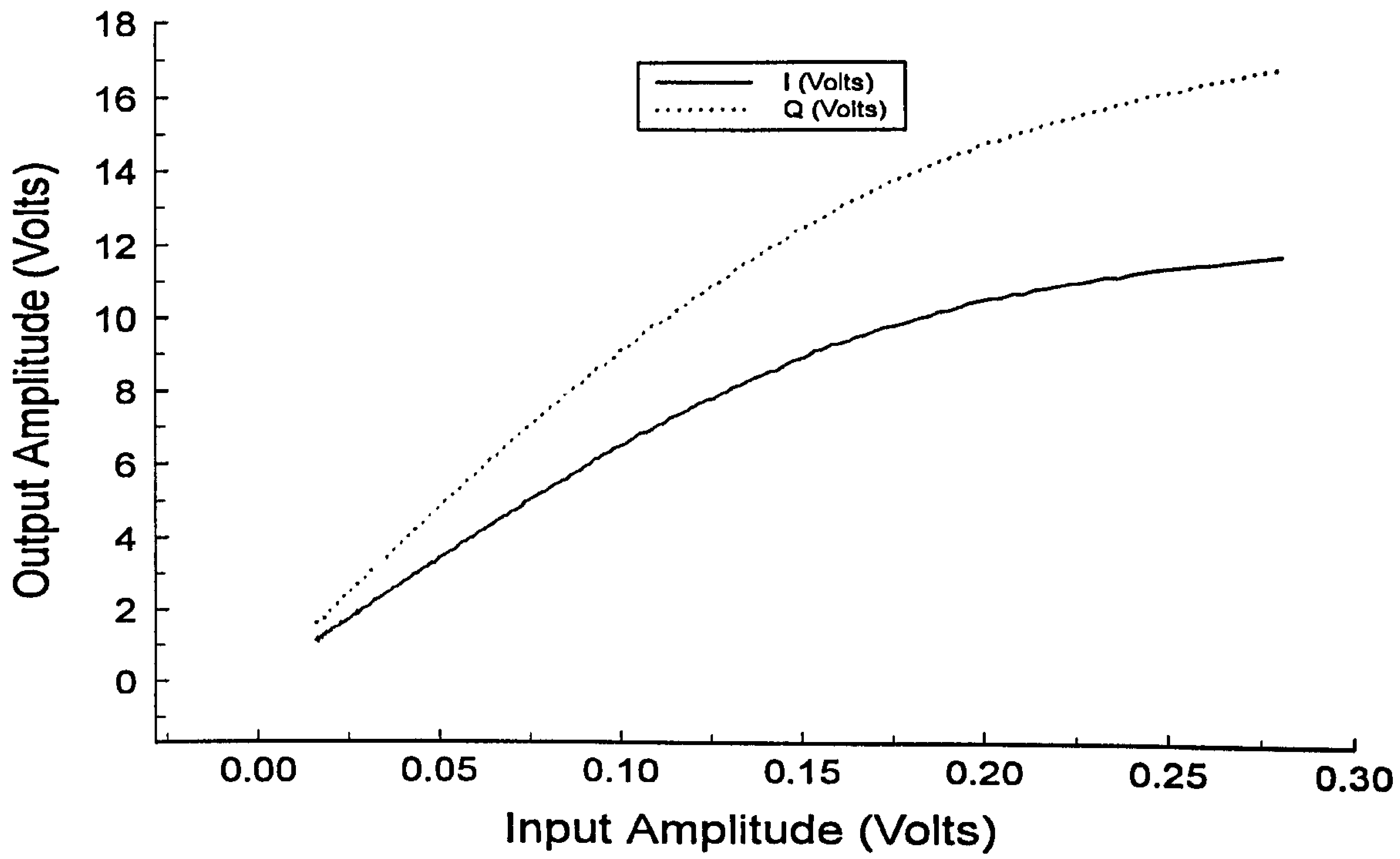

Figure 4.37 Amplifier I and Q Transfer Function

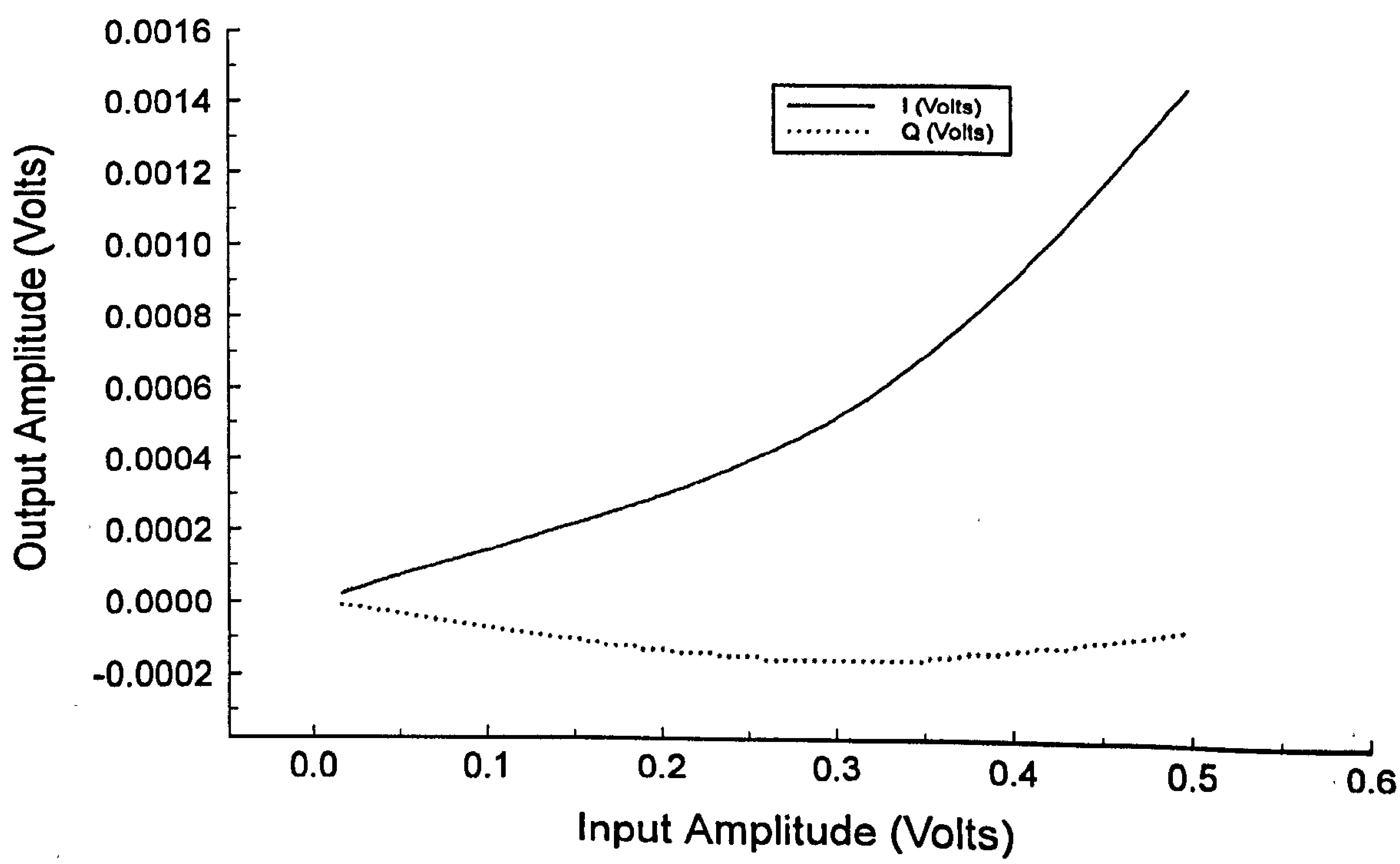

Figure 4.38 Cubic Predistorter I and Q Transfer Function

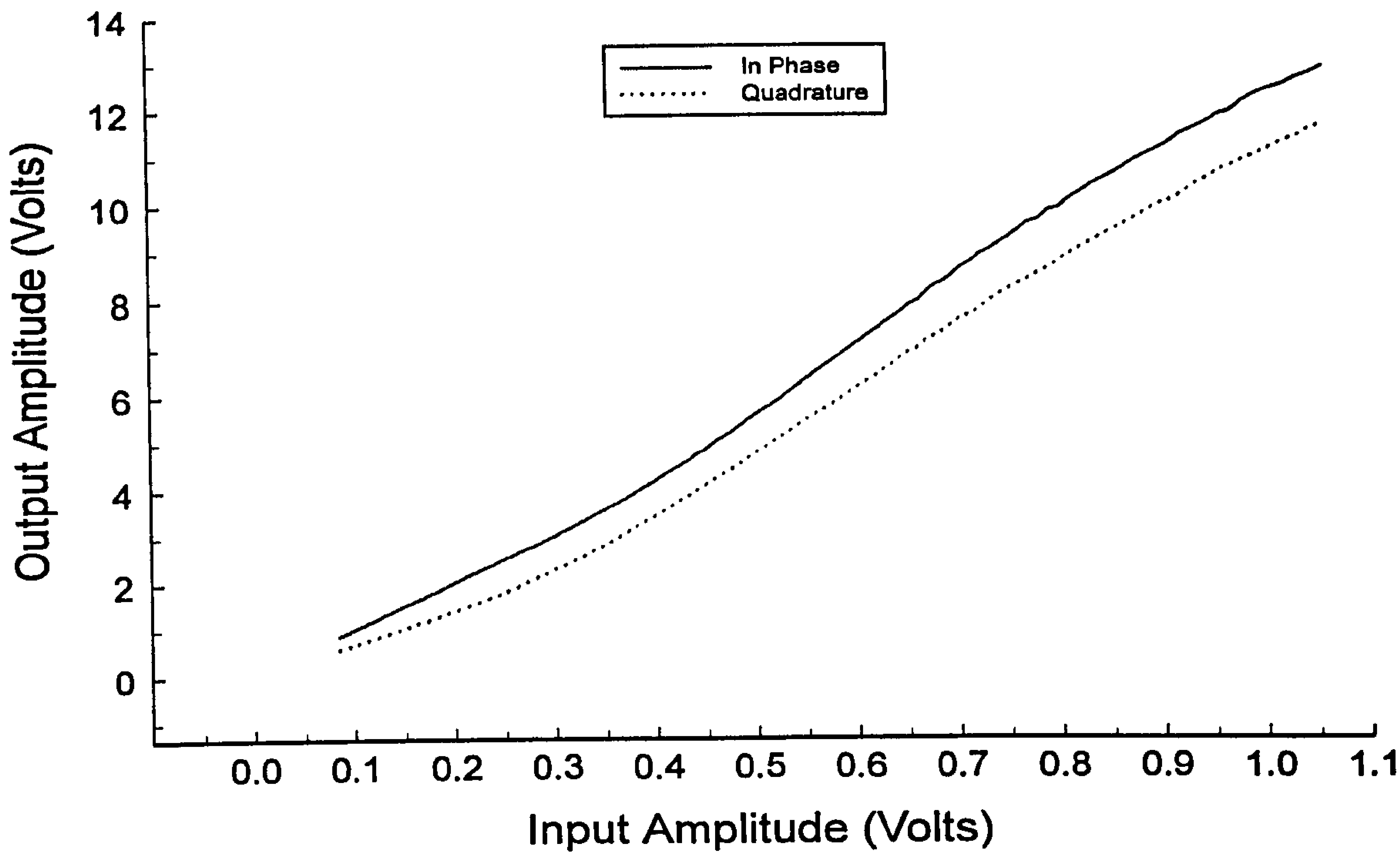

Figure 4.39 Cubic Predistorter plus Amplifier I and Q Transfer Function

Figures 4.37 and 4.38 show that the in phase components generated by the cubic predistortion system are the inverse of the components generated by the amplifier. The predistorter does not generate sufficiently negative quadrature components to cancel out the quadrature components generated by the amplifier. The net result of this is shown in figure 4.39, which shows how the systems in-phase response is greatly improved, but the system will still introduce quadrature distortion to any signal applied to it.

4.3.8 Alternative Forms of Cubic Polynomial Predistortion Element

The mixer-based form of polynomial element, which has been demonstrated thus far, has been shown to provide usable improvements in amplifier performance. The mixer based approach does however have the disadvantage of generating a characteristic which contains products of other orders apart from just the order which was required. The reason for this is that the mixers are not operating as pure multiplying elements. If a pure multiplier were used then potentially better levels of amplifier performance are possible. The analogue multiplier circuit has existed for many years [17]. This circuit may be used to generate a much purer form of polynomial characteristic. So using the same circuit architecture for the cubic element as before which is shown in figure 4.3. The new 3rd order predistortion element using Analog devices AD835 [18] analogue multiplier circuits was tested at a centre frequency of

100MHz with a tone spacing of 500kHz, the output spectrum for the modified cubic element is shown in figure 4.40.

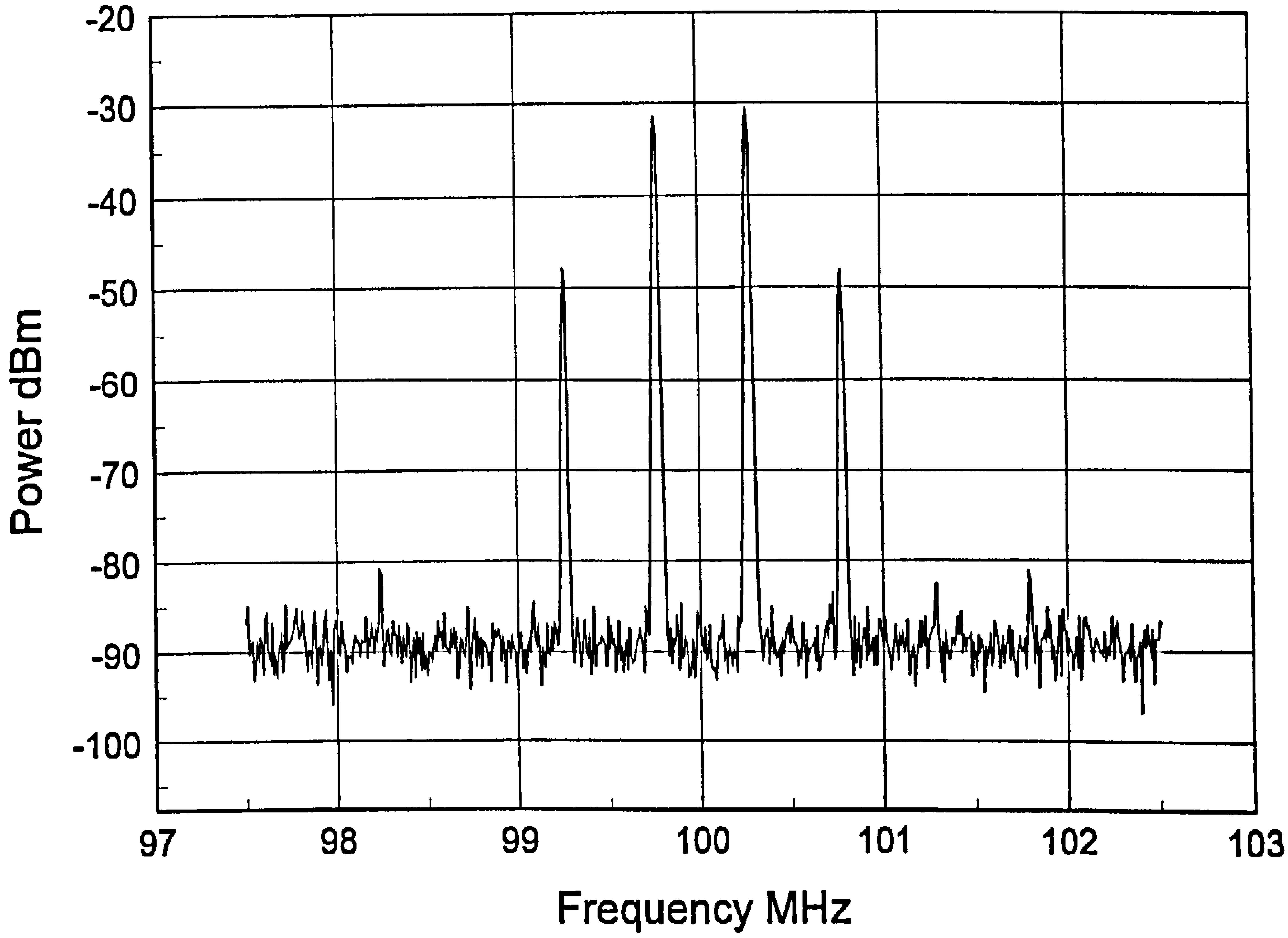

Figure 4.40 Output Spectrum for the Multiplier Based 3rd Order Predistortion Element

Figure 4.40 shows quite clearly that the output spectrum does indeed contain only the original tones plus the 3rd order IMP's with no additional products present.

This new element was then incorporated within the predistortion system shown in figure 4.5, the modified system was then connected to the amplifier. The amplifier was then tested to see the effect the modified elements would have on the performance of the predistortion system, tests were conducted at a centre frequency of 100MHz for a tone spacing of 500kHz. The amplifier two tone test is shown in figure 4.9 and the resulting predistorted output is shown in figure 4.41.

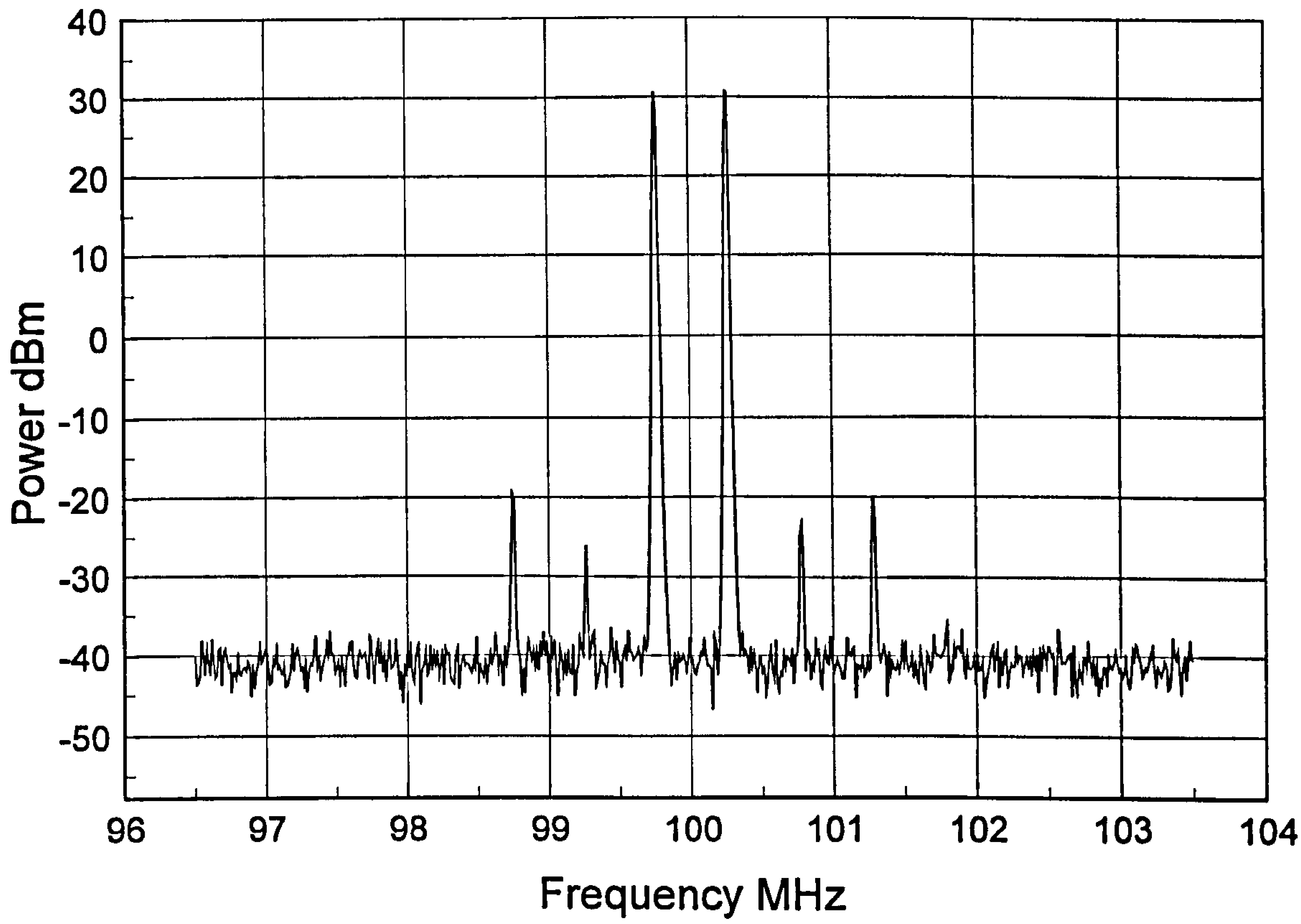

Figure 4.41 Amplifier Predistorted Output using Analogue Multipliers in the Cubic Predistortor

Figure 4.41 shows that greatly improved performance results may be achieved with use of analogue multipliers as the predistortion elements. The graph shows that there are no IMP's at greater than -50dBc. The 3rd order IMP's are at -53dBc which is a 23dB improvement over the unpredistorted case. These measurements seem to indicate that analogue multipliers are the preferred components for use in a polynomial predistortion system. However at present analogue multipliers are only available for use at frequencies of up to 500MHz [18], this makes their use for the RF predistortion of amplifiers at UHF impossible at the present time. Secondly the broadband performance of the analogue multiplier implementation of the predistorter is inferior to the mixer-based implementation.

4.3.9 Multiplier Predistorter Transfer Characteristics

The IMD performance of the analogue multiplier cubic predistorter is superior to the IMD performance of the mixer-based implementation. So it should follow that the transfer function of the multiplier-based system should show a more linear relationship than the results obtained by the mixer-based system. Investigations of the transfer functions of the predistorter and the combined system response were carried out these results are shown in figures 4.42 and 4.43.

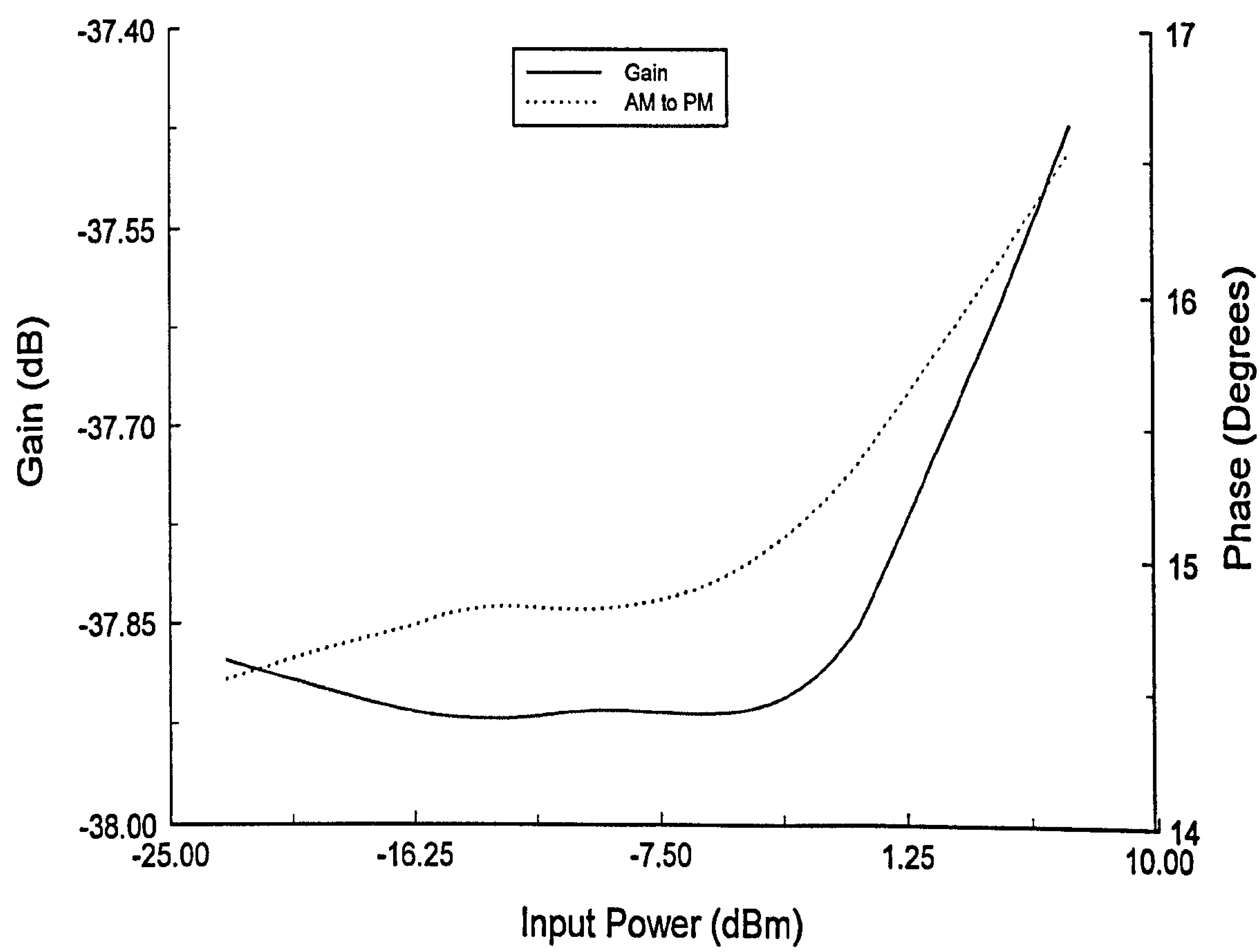

Figure 4.42 Multiplier Predistorter Gain and Phase Response

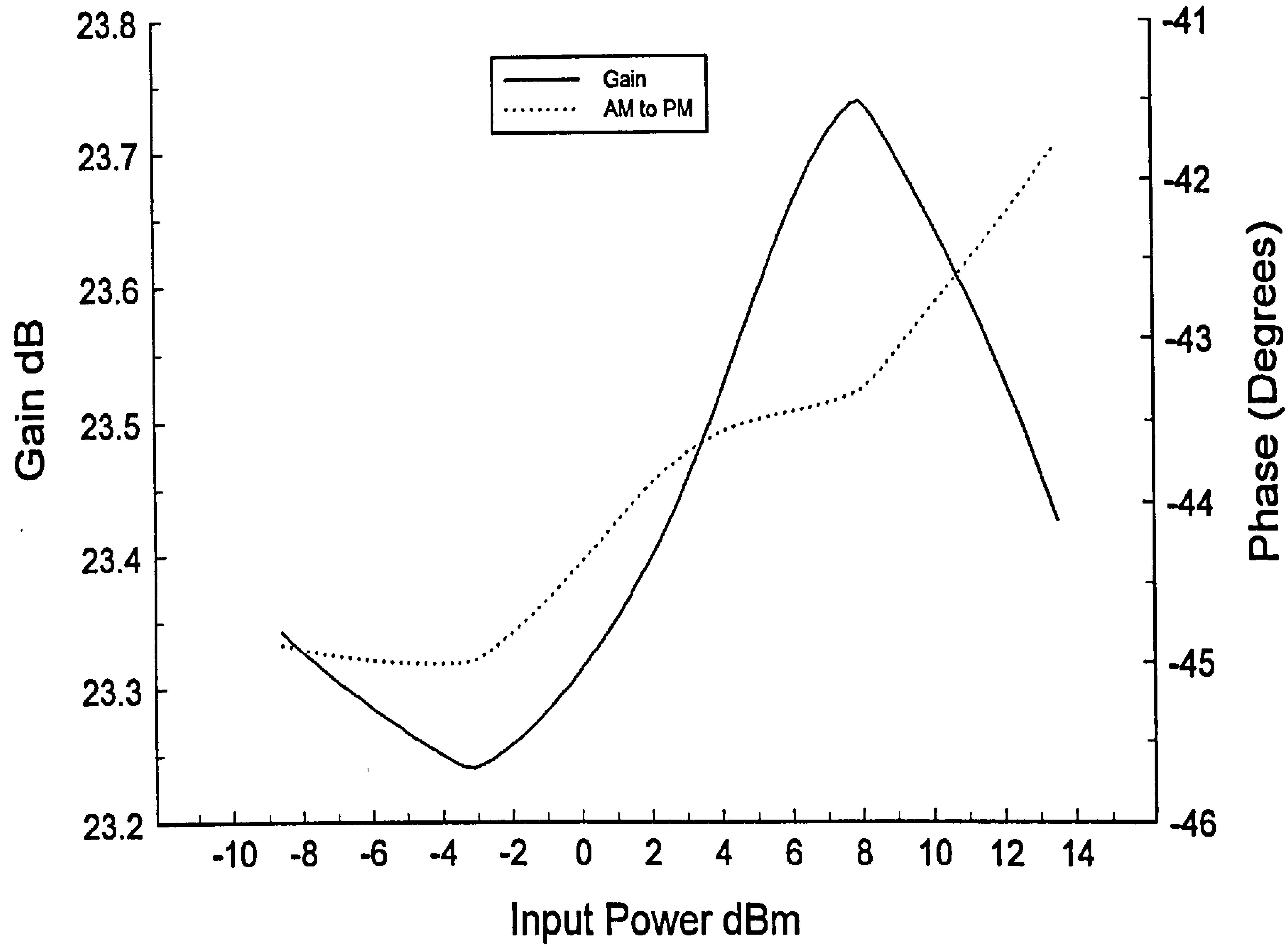

Figure 4.43 Multiplier Predistortor plus Amplifier Gain and Phase Response

The predistorter gain characteristic is similar in form to the characteristic generated by the mixer implementation. The predistorter phase characteristic is of a similar shape but shows a much lower level of amplitude change than the mixer based system, this is shown by figure 4.42. This means that a lower level of net error is present in the combined response shown in figure 4.43.

The predistorter transfer characteristic in figure 4.44 shows how the AM to AM characteristic is almost linear in form but the AM to PM characteristic introduces a phase change of approximately 2° over the whole power range. This is much lower than the mixer-based system, which exhibits 30° of AM to PM over its whole power range. This translates into an improved AM to AM and AM to PM composite characteristic which is shown in figure 4.45. Figure 4.46 shows that the in phase component increases at a much greater rate than the quadrature component.

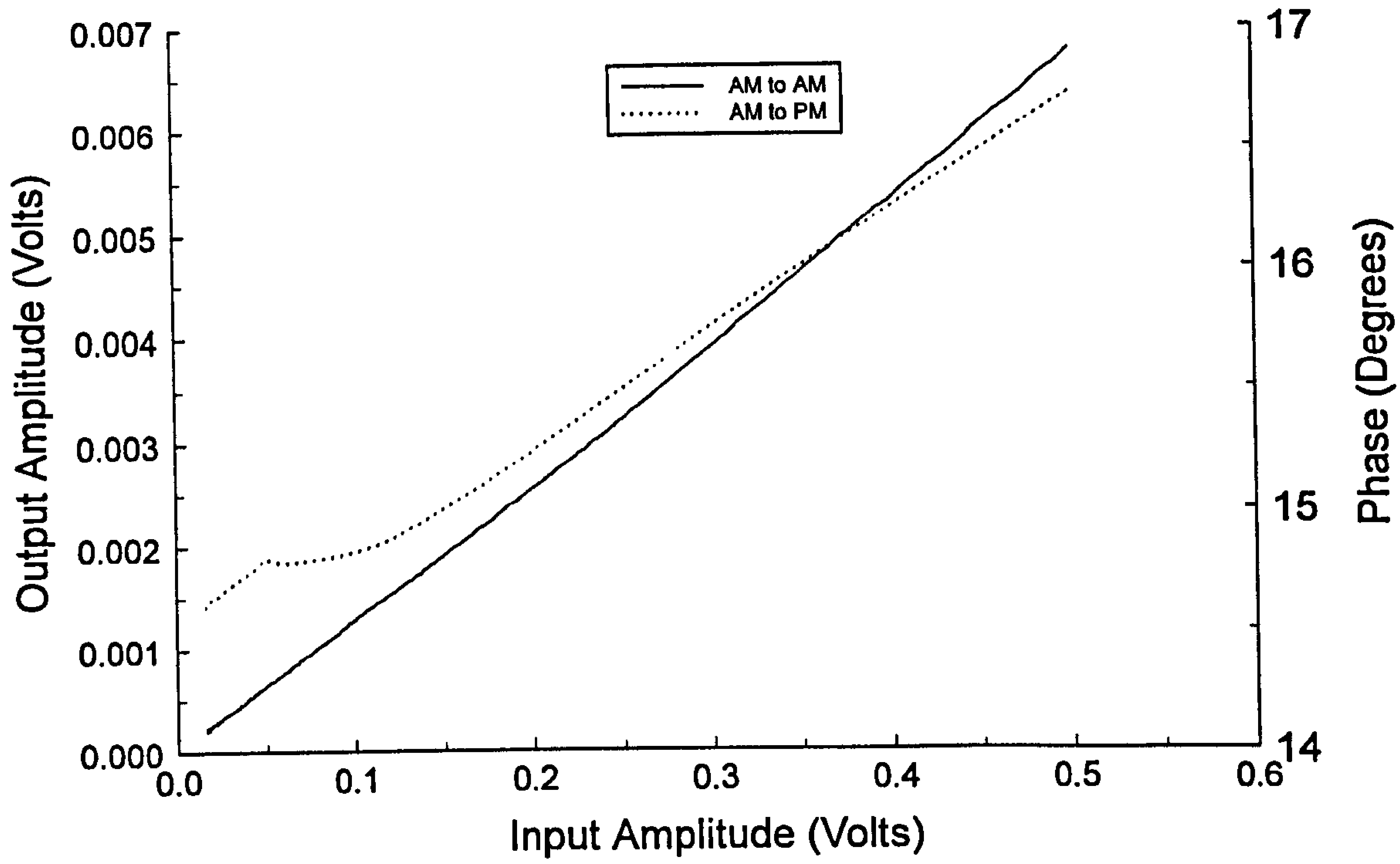

Figure 4.44 Multiplier Predistortor Voltage Transfer Function

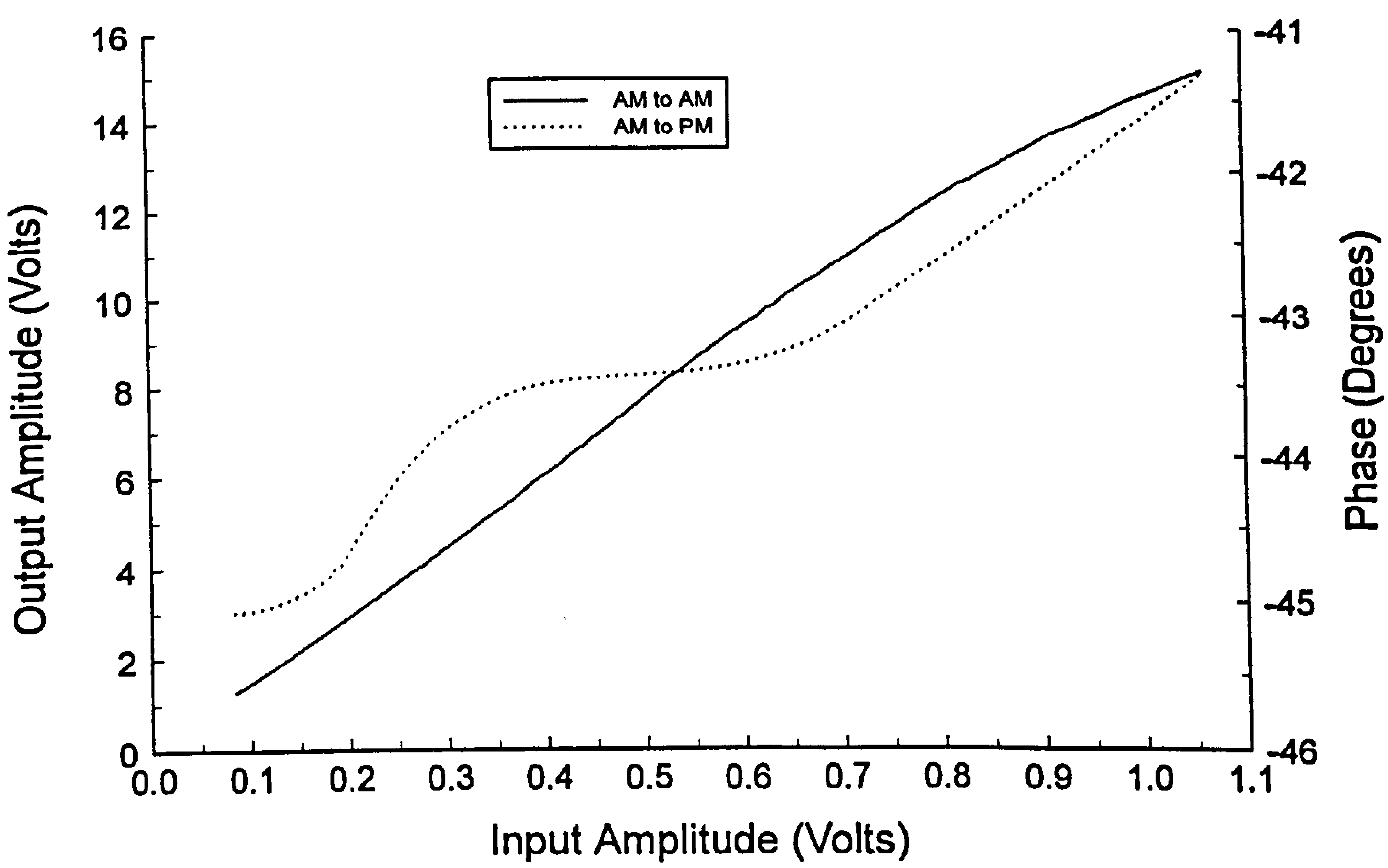

Figure 4.45 Multiplier Predisterter plus Amplifier Voltage Transfer Function

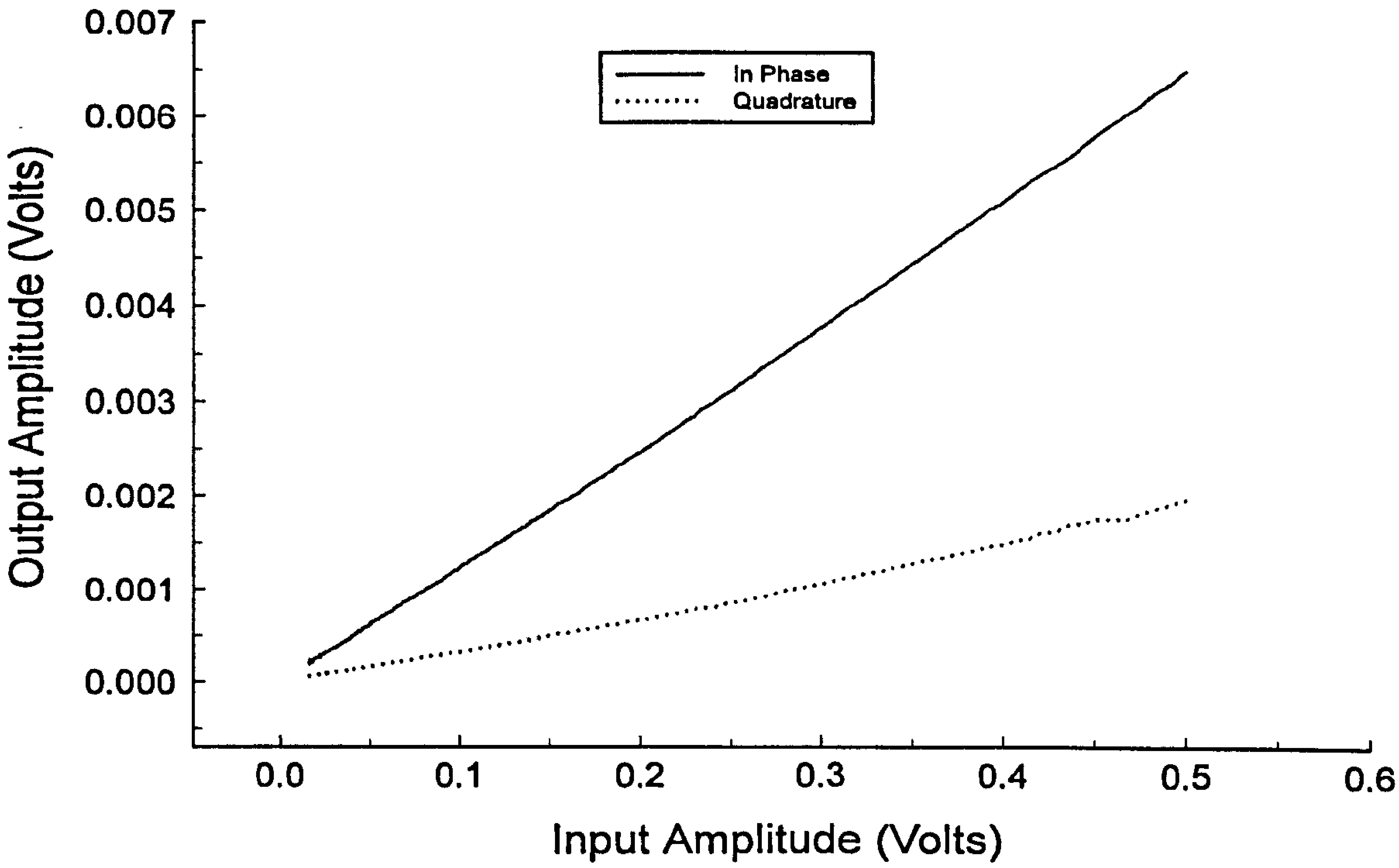

Figure 4.46 Multiplier Predistorter I and Q Transfer Function

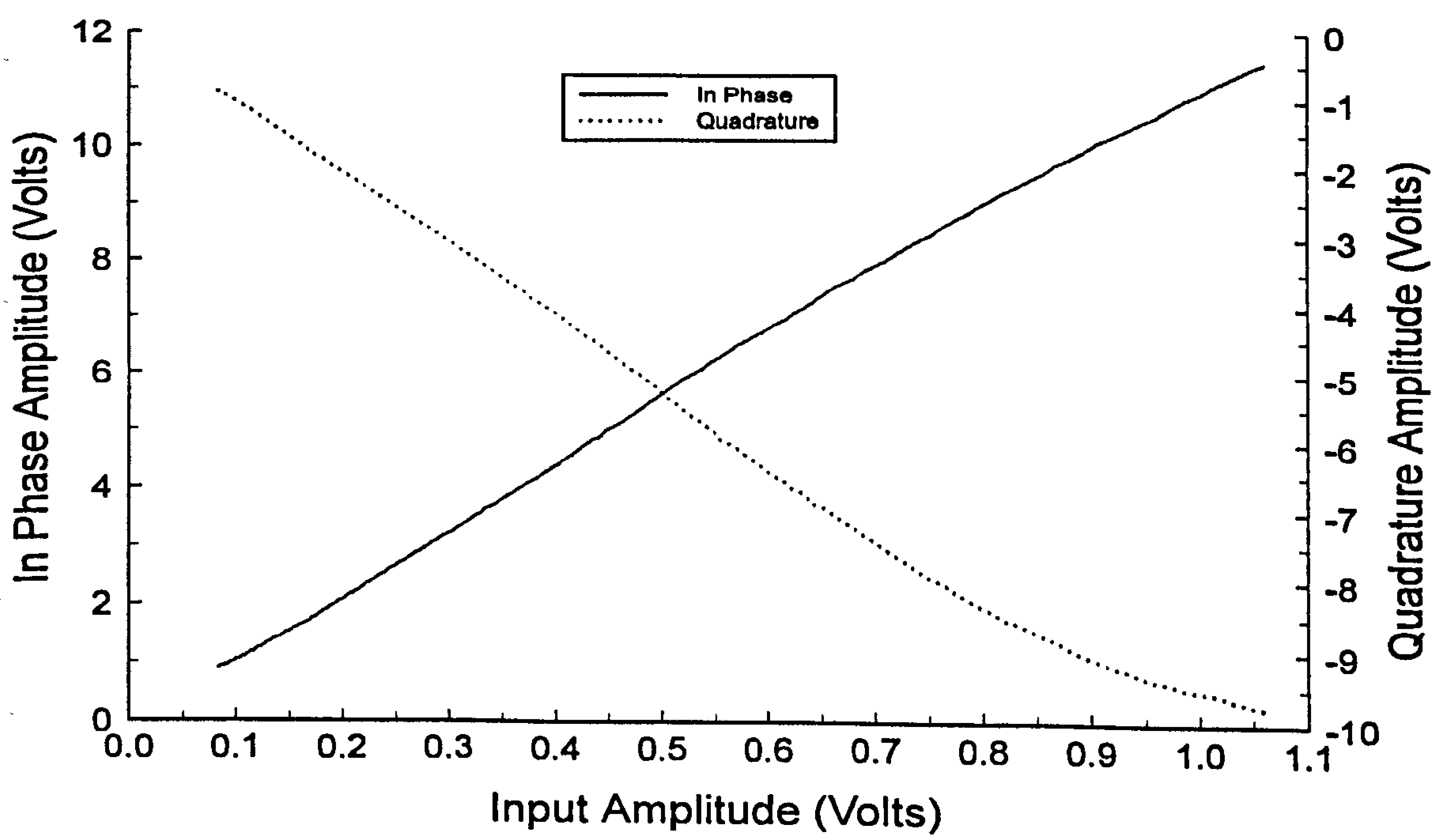

Figure 4.47 Multiplier Predistorter plus Amplifier I and Q Transfer Function

4.4 Summary

This chapter has introduced a new form of polynomial predistorter that has the capability of linearising an amplifier over a much greater bandwidth than has been reported previously [19]. This predistorter may be used to linearise an amplifier over a decade of frequency and at an instantaneous bandwidth of 180MHz. It has also been shown how the performance of the predistorter varies with gain and phase matching, wavelength error and amplifier back-off. Improvements in the overall system transfer characteristic are also shown within this chapter. These improvements in the transfer function allow the amplifier to be operated at the maximum output level that aids in the efficient operation of the amplifier.

References

[1] M. Gahderi, S. Kumar and D. E. Dodds, Adaptive Predistortion Lineariser Using Polynomial Functions, IEE Proceedings on Communications, volume 141, No. 2, pages 49 – 55, April, 1994.

[2] M. Gahderi, S. Kumar and D. E. Dodds, Fast Adaptive Polynomial I and Q Predistorter with Global Optimisation, IEE Proceedings on Communications, volume 143, No. 2, pages 78 – 86, April, 1996.

[3] Mohammad Ghaderi and Surinder Kumar, Power Amplifier Linearization using predistortion with Polynomial Functions, IEEE Vehicular Technology Conference, No. 43, pages 1125 – 1130, May 1993

[4] Michael Faulkner and Matts Johansson and Warren Yates, Error Sensitivity of Power Amplifiers Using Pre-Distortion, IEEE Vehicular Technology Conference, pages 451 – 456, May 1991.

[5] S. P. Stapleton and J. K. Cavers, Integrated DSP/RF Design for an MSAT Transmitter, Second International Mobile Satellite Conference, pages 302 – 307, June 1990.

[6] Toshio Nojima and Tohru Konno, Cuber Predistortion Linearizer for Relay Equipment in 800MHz Band Land Mobile System, IEEE Transactions on Vehicular Technology, volume VT34, No. 4, pages 169 – 177, November 1985.

[7] Rodrigo Cordeiro Tupynamba and Edmar Camargo, MESFET Nonlinearities applied to Predistortion Linearizer Design, IEEE MTT-S, pages 955 – 958, June 1992.

[8] John A. Steck and David Pham, A New TWT Linearizer for Satellite Communications Transmitters, Microwave Systems News and Communications Technology, pages 28 – 43, June 1987.

[9] Shigenobu Aihara, Toshio Nishiumi, Yoshinori Fujiki and Sachiro Fukuda, GaAs FET Power Amplifiers as Substitutes for TWT Amplifiers in A Multilevel QAM Digital Radio System, ICC, volume 1, pages 6 – 10, June 1987.

[10] Hideo Ashida, Yoshiaki Suzuki, Isamu Umino and Norio Tozawa, C Band 100 Watts GaAs FET Amplifier for Digital Radio, ICC, volume 1, pages 23 – 27, June 1987.

[11] D. Chakrabarty and J. M. Kappes, Some Aspects of Wideband H. P. A Linearizer Testing, IEEE National Telesystems Conference, pages 311 – 316, November 1982.

[12] Mahesh Kumar and James C. Whartenby, TWTA Linearizer using Dual Gate MESFET, IEEE MTT-S, pages 314 – 315, June 1984.

[13] Mahesh Kumar, James C. Whattenby, Herbert J. Wolkstein, Predistortion Linearizer Using GaAs Dual Gate MESFET for TWTA and SSPA used in Satellite Transponders, IEEE Transactions on Microwave Theory and Techniques, volume 33, No. 12, pages 1479 – 1488, December 1985.

[14] Paul H. Young, Electronic Communication Techniques, Third Edition, Macmillan Publishing Company, New York, 1994.

[15] Kevin Morris and Peter Kenington, Power Amplifier Linearisation Using Predistortion Techniques, IEE Colloquium on RF and Microwave Components for Communications Systems, pp. 6/1 – 6/6, April 1997.

[16] Mini Circuits RF/IF designers handbook, 1997

[17] B. Gilbert, A Precise Four-Quadrant Multiplier with Subnanosecond Response, IEEE Journal of Solid-State Circuits, volume SC3, No. 4, pages 365 – 373, December 1968.

[18] Analog Devices IC Handbook 1992.

[19] Kevin Morris and Peter Kenington, A Broadband Linear Power Amplifier for Software Radio Applications. In Proceedings of the 48th IEEE Vehicular Technology Conference, pp. 2150 – 2154, Ottawa May 1998.

Chapter 5
Higher Order Polynomial Predistortion

This chapter discusses the application of higher order predistortion techniques for the linearisation of R.F. power amplifiers

5 Higher Order Polynomial Predistortion

5.1 Introduction

Chapter 4 has shown that analogue predistortion may be used for the broadband linearisation of RF power amplifiers, the possibility of extending this for higher order systems has also been discussed. Higher order polynomial predistortion systems have received little attention in the literature, mainly due to the fact that linearisation has been generally attempted on moderately non-linear amplifiers with mildly compressive characteristics. Amplifiers with purely compressive characteristics such as class A and class AB amplifiers are generally unsuitable for use in mobile terminals due to their poor efficiency. Also it is desirable to improve the efficiency of the base station terminal in order to reduce size and power dissipation. To achieve these aims it is necessary to improve the linearity of the more efficient classes of amplifier. More efficient amplifiers do not have characteristics that can even be approximately [1, 2] modelled using single order non-linearities. This chapter details various methods of generating more complex polynomial characteristics.

5.2 Higher Order Predistorters

It has been stated that the polynomial predistortion techniques employed in chapter 4 may be extended to higher orders by the use of additional multiplication blocks. The following sections detail investigations into higher order systems initially concentrating purely on fifth order systems and then covering combined third and fifth order systems.

5.3 Quintic Predistortion Systems

An implementation for a quintic predistorter is shown in figure 5.1.

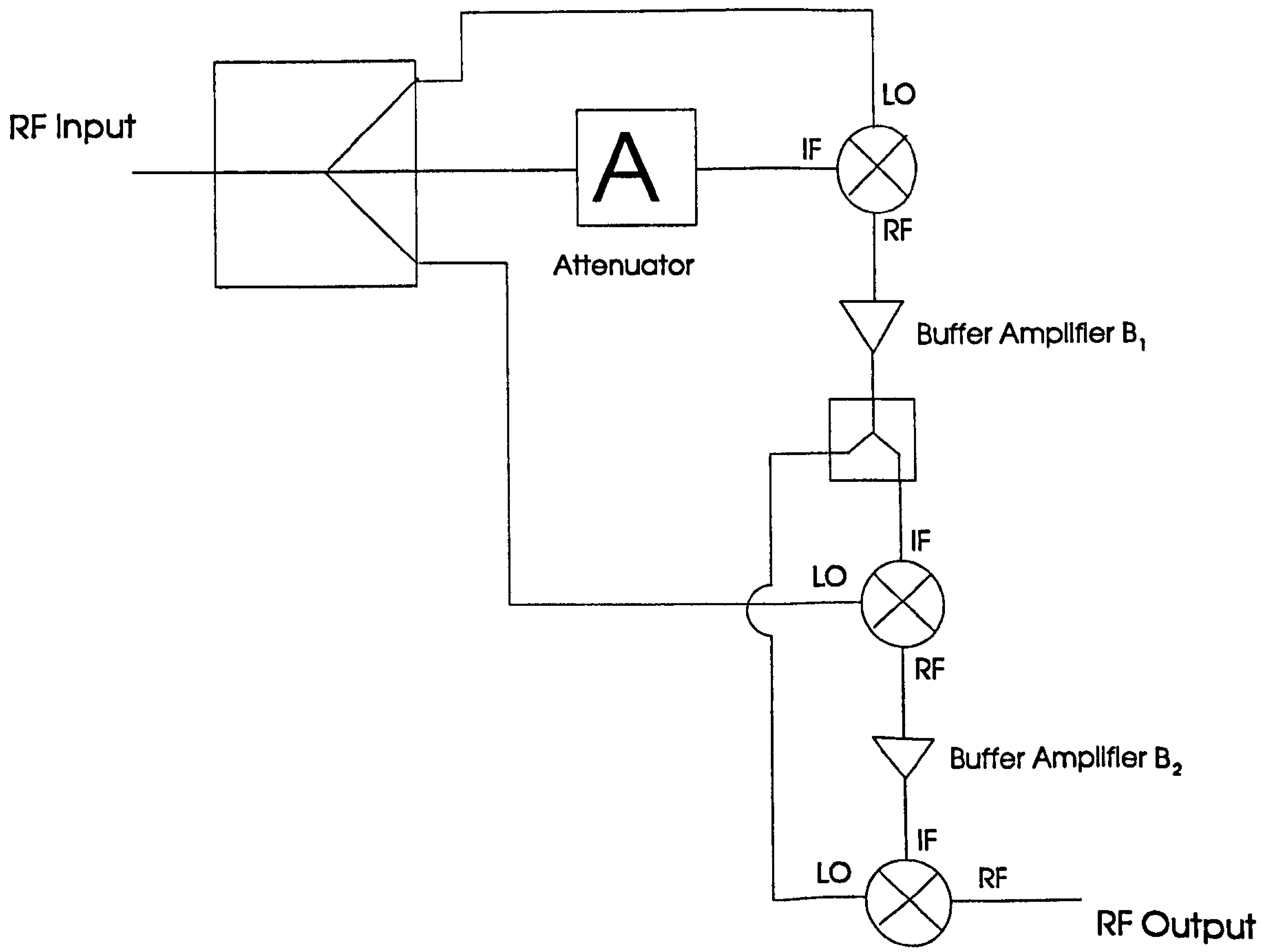

Figure 5.1 The Quintic Predistorter Element

The circuit operates in the following manner, the RF signal is applied to the three-way splitter where it is divided equally and applied to the first IF port via an attenuator, which ensures that the IF port is not overdriven. The signal is also applied to the first LO. and to the second LO. The first mixer provides the square of the input function at the R.F. port. This signal is then fed via a buffer amplifier, which supplies sufficient gain to compensate for the losses due to the mixing process, to a second power splitter where the signal is applied to the second mixer IF and the third mixer LO. The combination of the original signal at the LO and the squared signal at the IF produces a cubic signal at the output of the second mixers RF port. This signal may be used for cubic predistortion purposes or as the source signal for higher order products. The RF output of the second mixer is applied via a second buffer amplifier, which provides sufficient gain to compensate for the losses associated with the mixing process, to the IF input of the third mixer. This signal is combined with the second order signal at the mixer LO to produce the quintic output at the third mixer's R.F. output.

The quintic element was tested at 220MHz with a tone spacing of 500kHz and a tone power of 9dBm per tone, the results of these measurements are shown in figure 5.2.

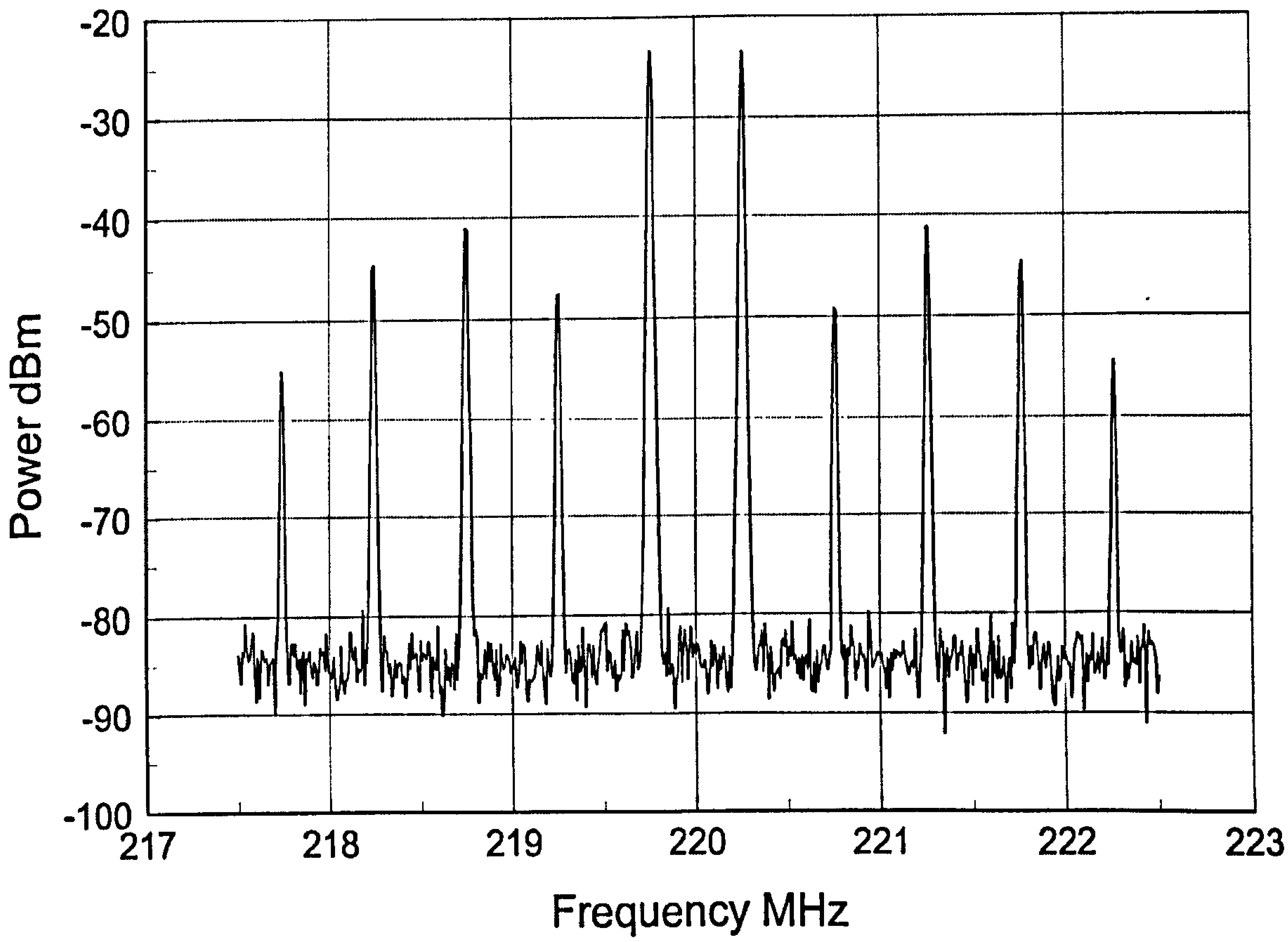

Figure 5.2 5th Order Element Output at 220MHz with 500kHz Tone Spacing

It may be observed that the 5th order products are at –15dBc whilst the 3rd order products are at –23dBc. The power of the 7th and 9th order products is considerable when compared with the 5th order products and so the element is not a pure producer of 5th order products.

This quintic element was incorporated within the predistortion system shown in figure 5.3.

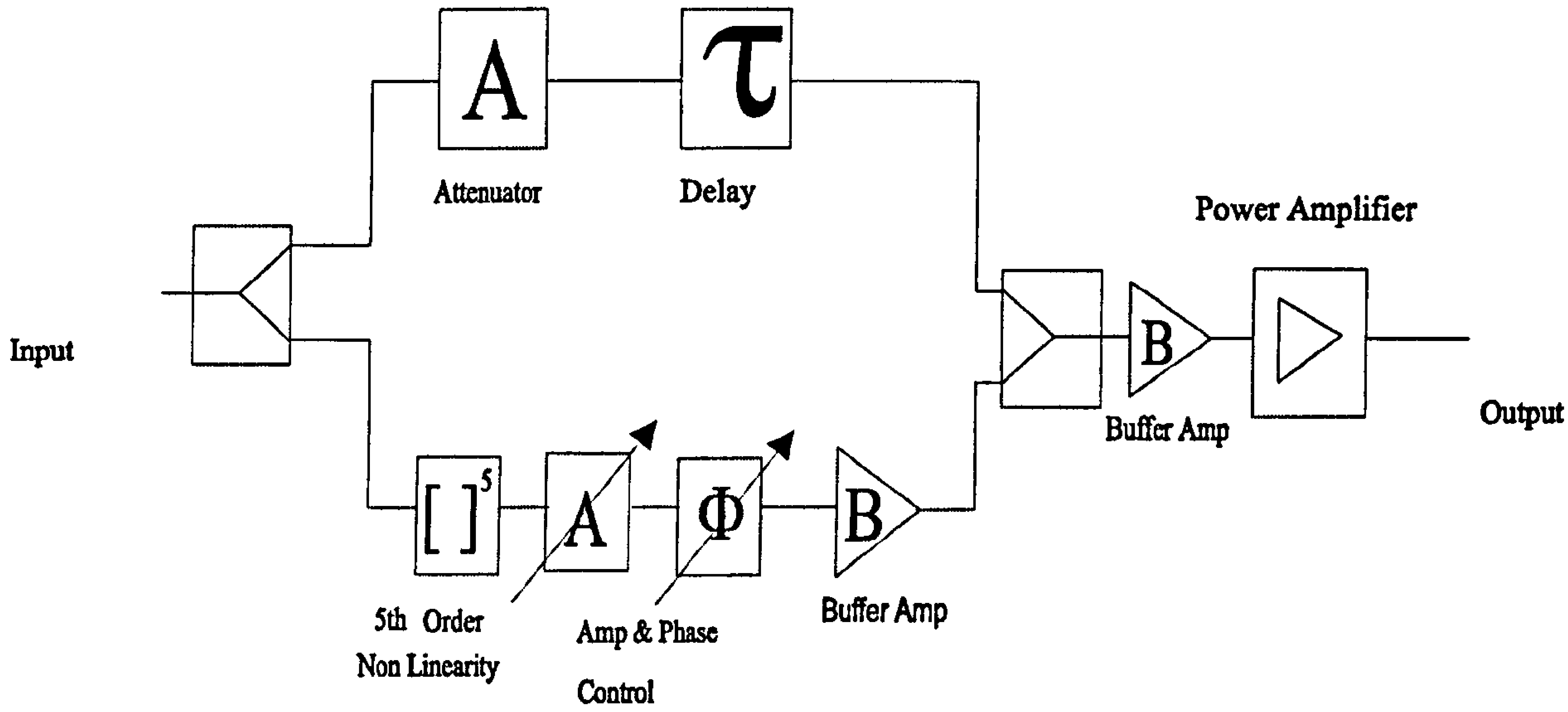

Figure 5.3 5th Order Predistortion System

The predistortion system is very similar in form to the system used in chapter 4 for cubic predistortion purposes. The main modification being the inclusion of the 5th order element in the position previously occupied by the cubic element. The attenuator was set to 14dB and the buffer amplifier was a MAR 8 amplifier from Mini Circuits that had been optimised for operation at 220MHz. The power amplifier used in chapter 4 was first two-tone tested and then predistorted using the predistortion system, these plots are shown in figure 5.4 and figure 5.5 respectively. The plots show that the 5th order products have been cancelled by 15dB with the 3rd order products being cancelled by 2dB, which may be seen in figure 5.5.

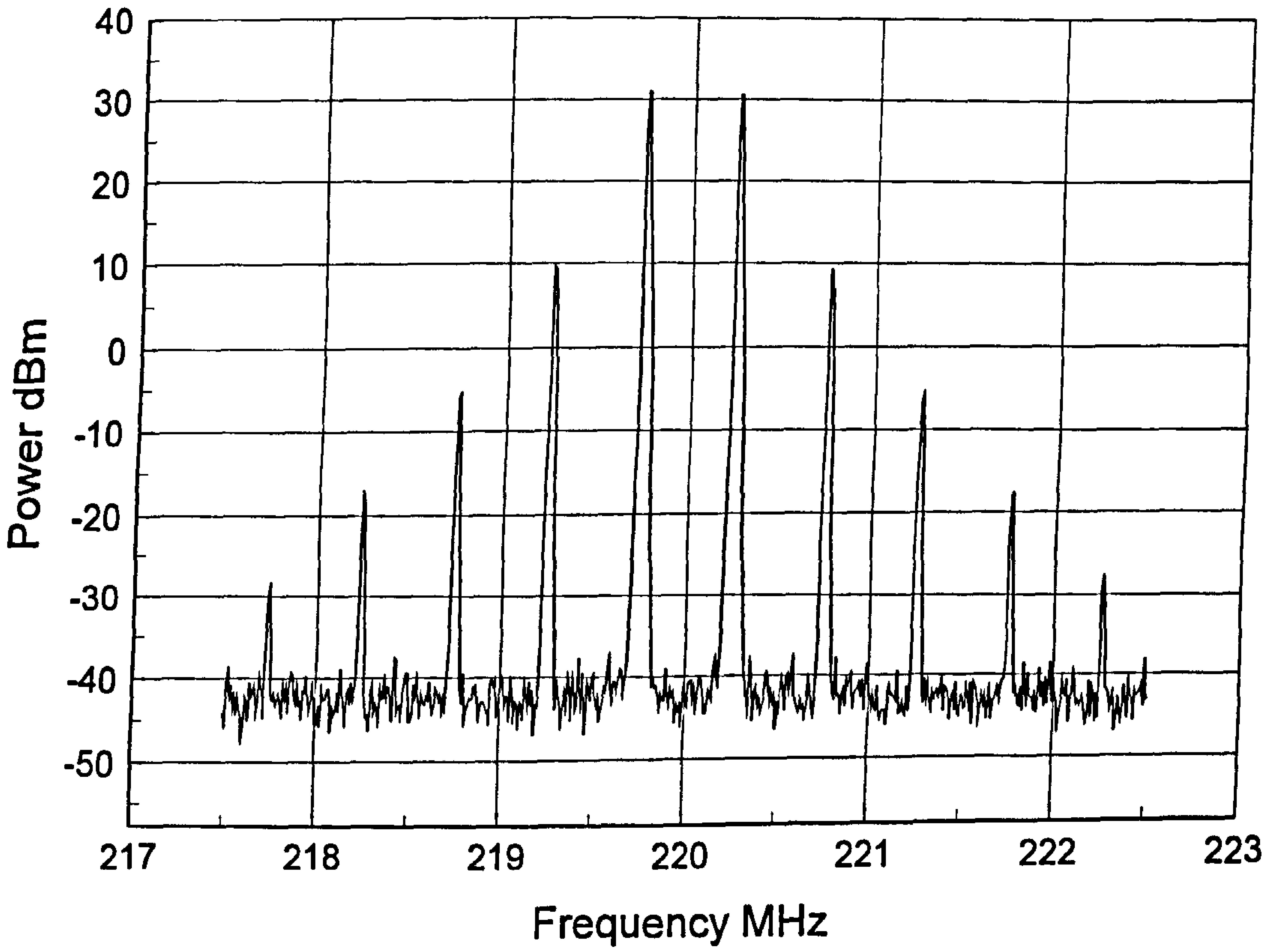

Figure 5.4 Amplifier Two Tone Test at 220MHz with 500kHz

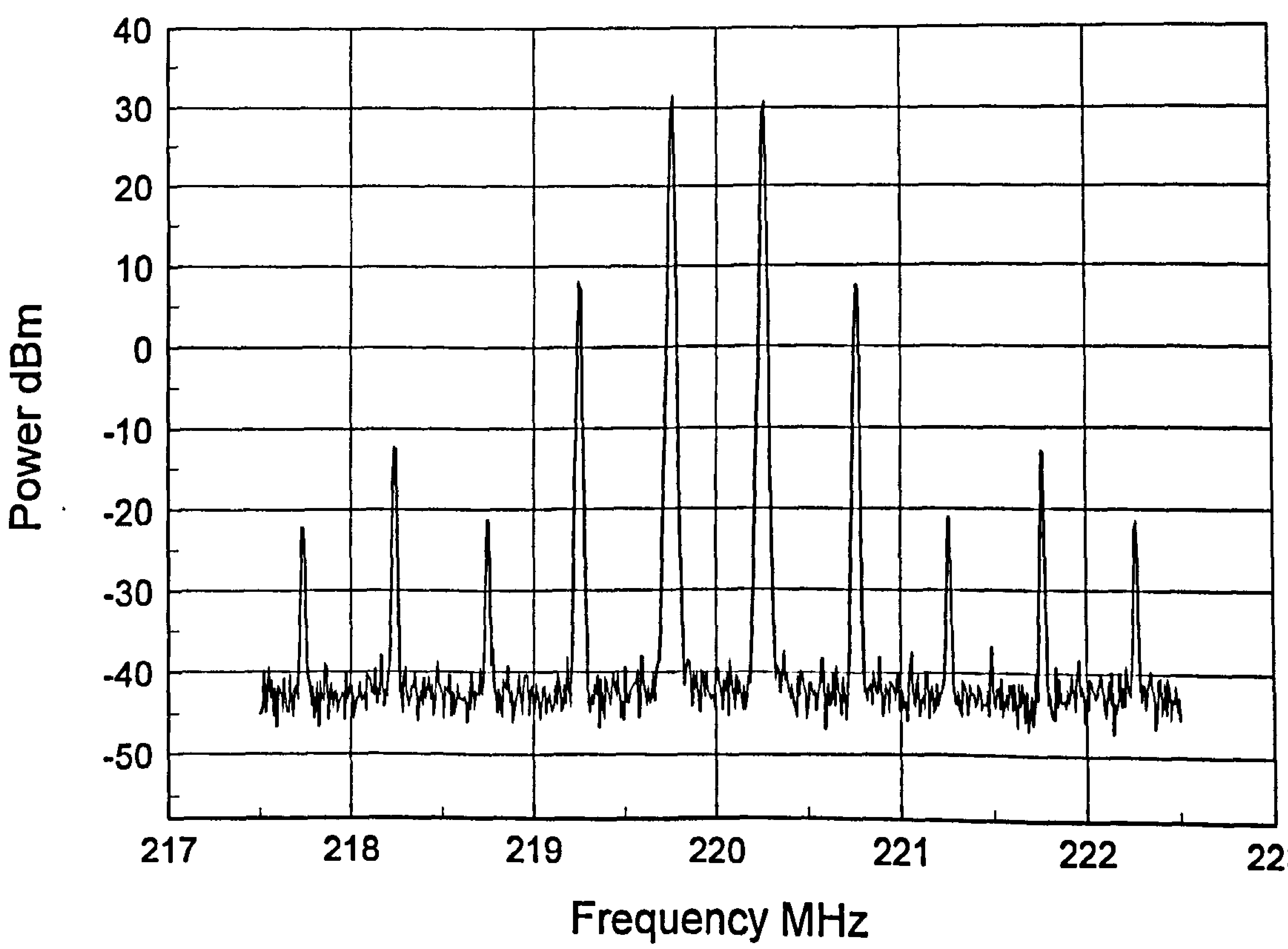

Figure 5.5 Amplifier Output After 5th Order Predistortion

5.3.1 Effect of Delay Error

It would be desirable to operate the quintic predistortion system over a bandwidth which is comparable with that achievable with the cubic predistortion system of chapter 4. The delay τ was set to optimise the performance of the quintic predistorter by making it equal to the delay through the predistorted path. The delay was then adjusted to measure the degradation in performance due to delay error, the results of these measurements may be seen in figure 5.6.

For errors in wavelength of +/- 0.1λ the performance has been degraded by 6dB, which results in a performance improvement over the unpredistorted case of 3dB. With an error of +/- 0.05λ the performance has been degraded by 3dB, which results in a performance improvement of 6dB over the unpredistorted case. So the quintic system is as tolerant of error as the previously developed cubic system.

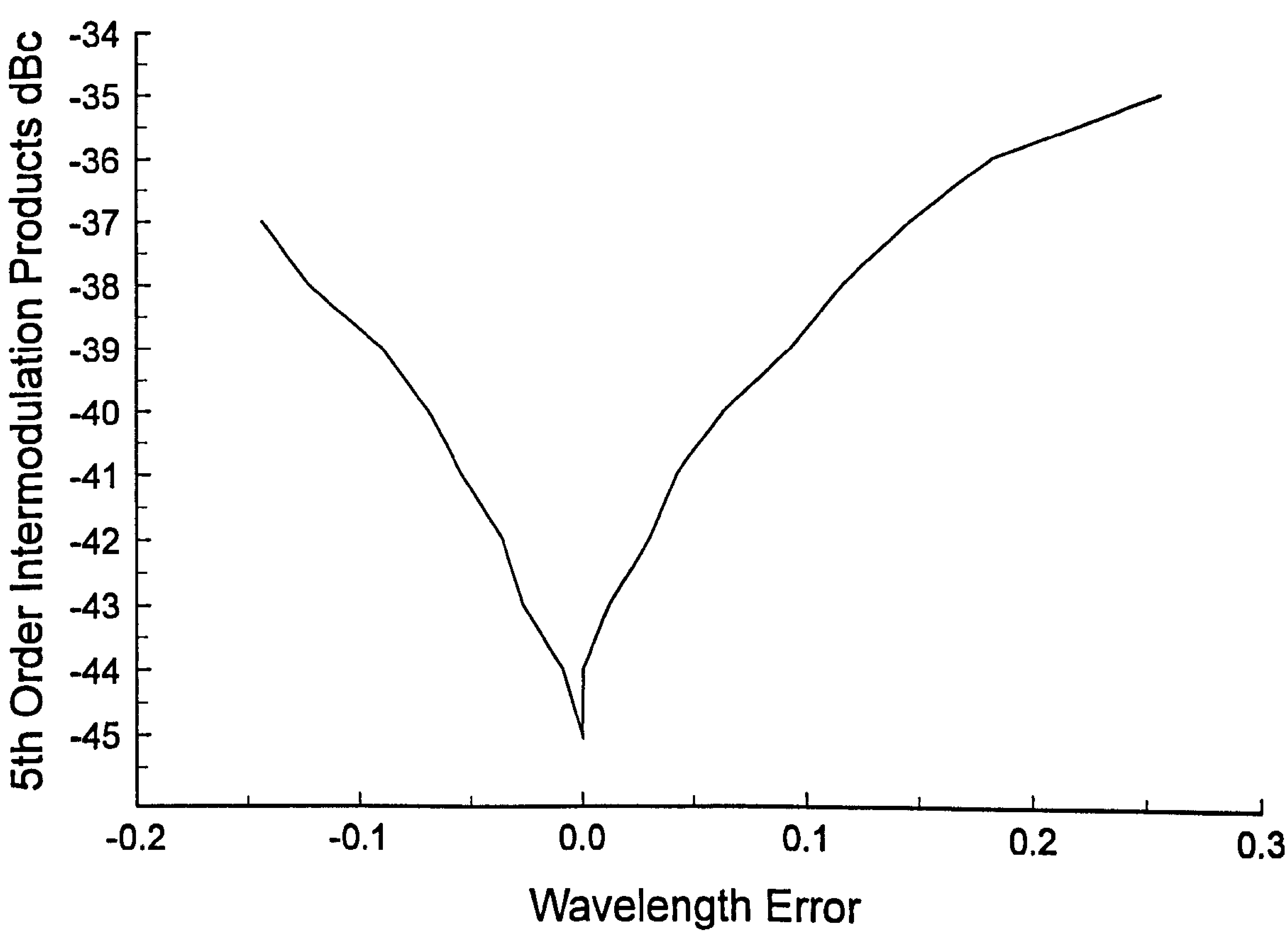

Figure 5.6 Effect of Wavelength Error on 5^{th} Order Intermodulation Products

5.3.2 Broadband Performance

Quintic predistortion relies on the generation of sum and difference products that are combinations of the third and second harmonic zones. For example the lower 5th order IMP is:

$$Frequency_{IMP5} = 3f_1 - 2f_2 \tag{5.1}$$

This implies that the predistortion element must be capable of dealing with frequencies of up to five times the carrier frequency (these IMP's are generated in the 5th harmonic zone). So for satisfactory performance the mixers within the quintic element must be capable of translating frequencies of up to 1100MHz (assuming a 220MHz carrier frequency). Mixers were chosen [3] which would operate at up to five times the carrier frequency.

The circuit in figure 5.3 was optimised to operate in a broadband mode. The amplifier used in previous measurements was then tested at 220MHz with a tone spacing of 20MHz. The measurements of the amplifier two tone test performance and predistorted output are shown in figures 5.7 and 5.8 respectively. If figures 5.7 and 5.8 are compared the following observations maybe made; the 3rd order IMP's are unchanged, the 5th order IMP's have been reduced by 5dB when compared with the unpredistorted case and the 7th order IMP's have also been reduced by 5dB. It is possible therefore to reduce the 5th order IMP's, however the 3rd order IMP's are unaffected.

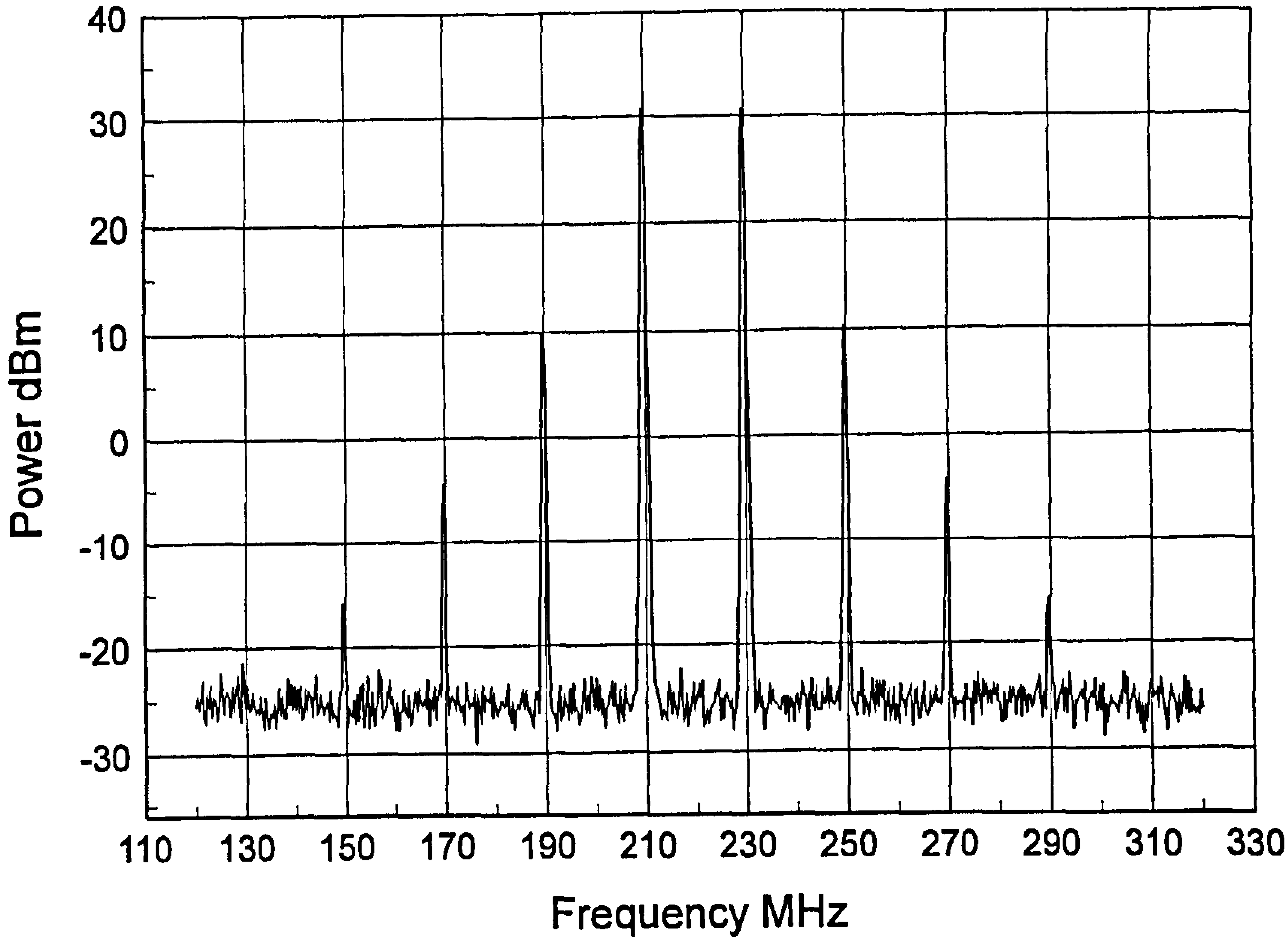

Figure 5.7 Two Tone Test at 220MHz with 20MHz Tone Spacing

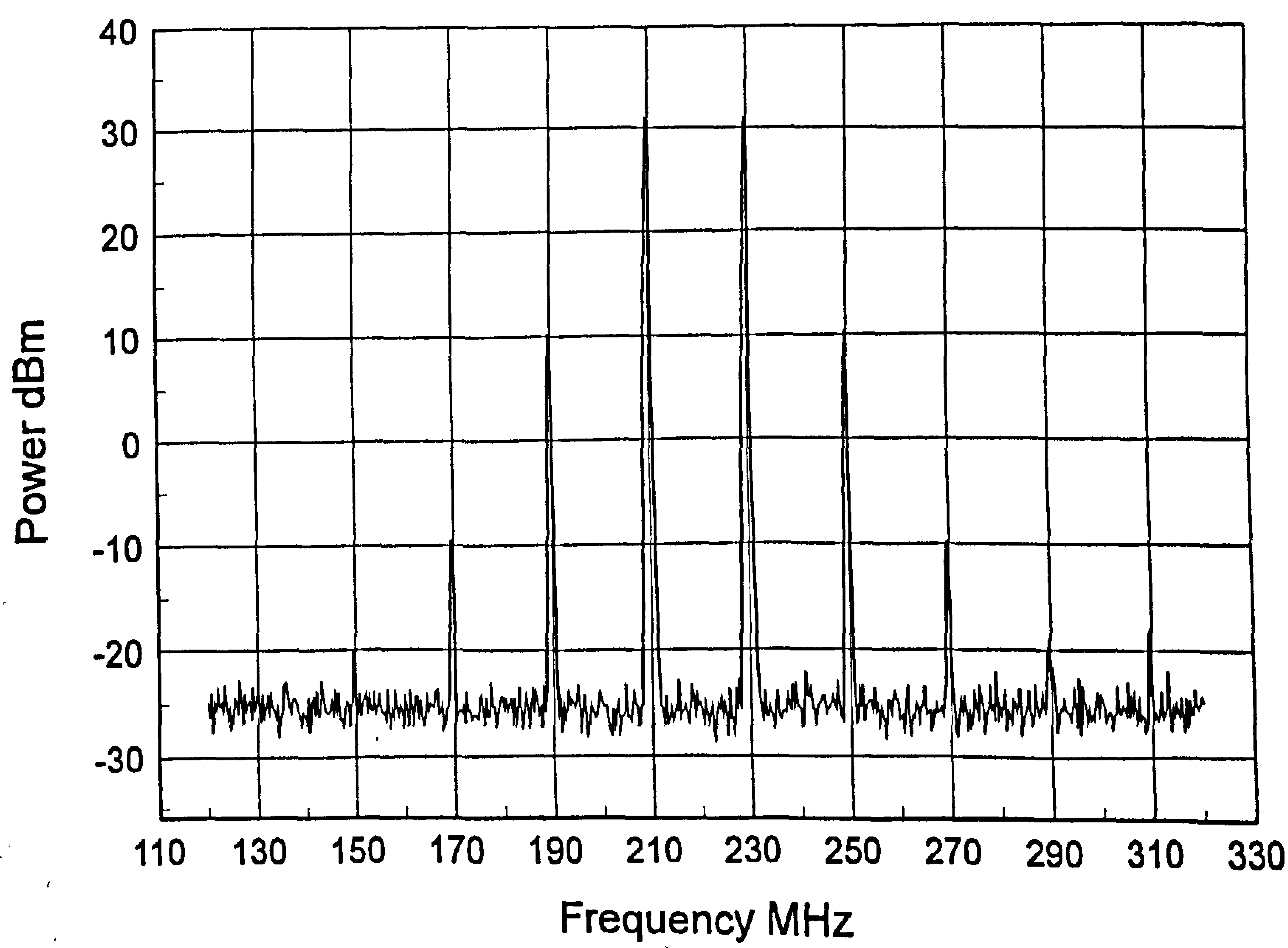

Figure 5.8 Predistorted Amplifier Output after 5th Order Predistortion at 20MHz Tone Spacing

5.3.3 Effect of Back-off on 5th Order Intermodulation Distortion

The intermodulation performance of class A and class AB amplifiers will change dependent on the input power level. So it is necessary for any predistortion system to be able to still provide improvements in performance under changing power conditions. For maximum efficiency it is desirable to drive the amplifier at its peak output power. But in practice an amplifiers output level will vary under varying load conditions. So the effect of back-off on intermodulation performance has been investigated for the quintic predistorter.

The amplifier's 5th order intermodulation product back-off performance was measured for three cases; with the amplifier lone, for the amplifier with the predistortion system fitted and optimised for every power level and finally an initial set up was carried out at the 1dB compression point and then the power level was adjusted without any further adjustment being made. The intermodulation product value was plotted for all three cases, the results are shown in figure 5.9.

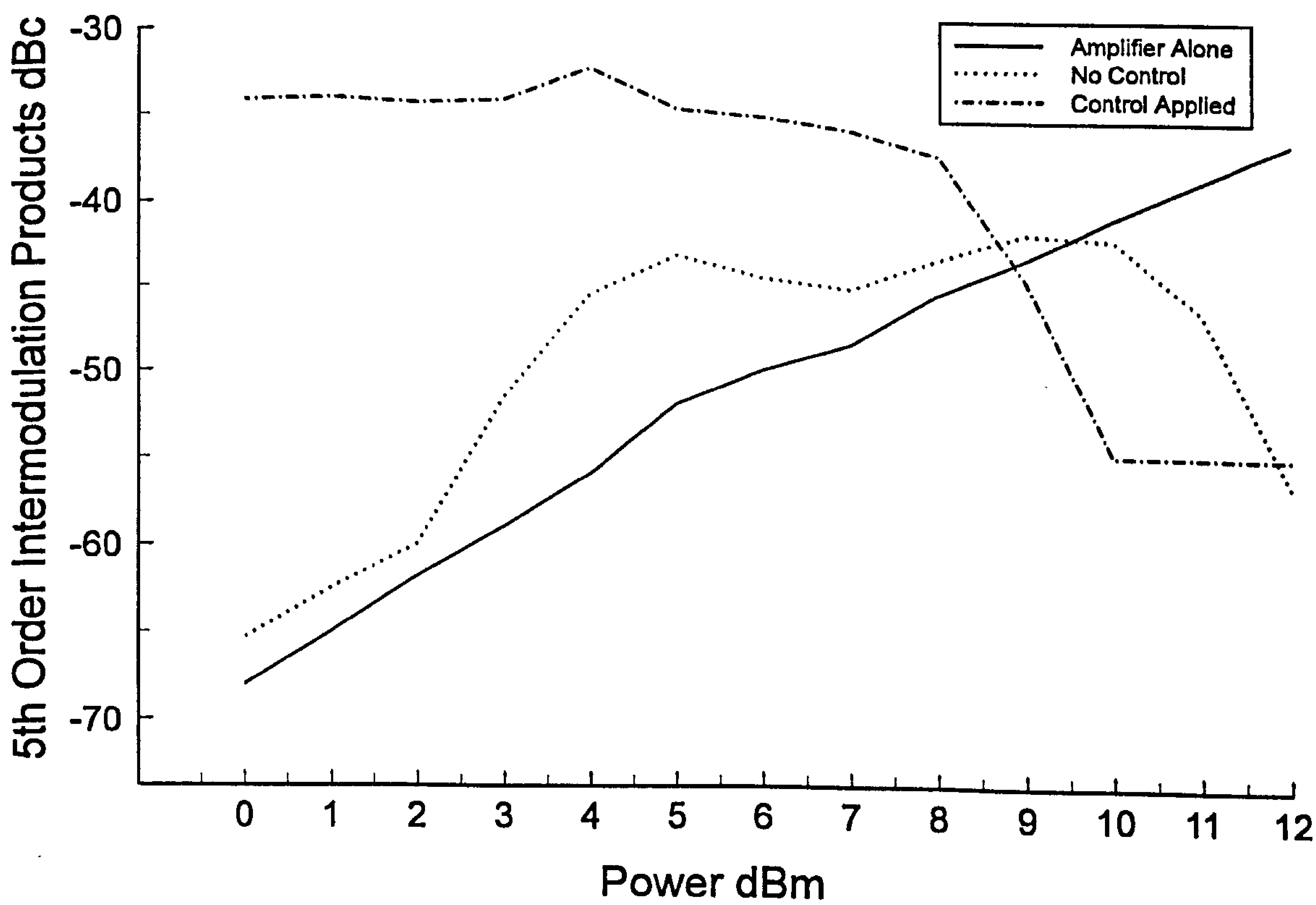

Figure 5.9 Effect of Back-off on 5th Order Intermodulation Performance

Figure 5.9 shows that when the unpredistorted and the predistorted case are compared at 12dBm input power (the 1dB compression point) the IMP's have been reduced by 20dB. It can be seen that useful reductions in 5th order IMP's are possible using the predistorter until the input power has fallen to 9dBm. If the input power is reduced further the IMP performance is worse than the amplifier with no predistortion applied. This effect is due to the predistortion system introducing an excess of IMP cancellation that results in degradation in amplifier performance.

5.3.4 Discussion

The quintic predistortion system can provide useful reductions in 5^{th} order IMP's. In general however the 3^{rd} order products will always dominate the performance of any amplifier system, so it is desirable to cancel both 3^{rd} order and 5^{th} order products simultaneously. The next section introduces a combined 3^{rd} and 5^{th} order predistortion system.

5.4 3^{rd} and 5^{th} Order Predistortion System

Section 5.3 discussed the development of a quintic predistortion system, this system has the disadvantage of only correcting for 5^{th} order non-linearity effects. An effective amplifier linearisation system will have to correct for at least 3^{rd} and 5^{th} order non-linearity. This section deals with the development of a system to correct for 3^{rd} and 5^{th} order non-linearity effects.

5.4.1 3^{rd} and 5^{th} Order Predistortion Element

It is possible to generate any order of non-linearity desired by using the mixer-based approach of chapter 4. It has been shown that it is possible to generate a 5^{th} order non-linearity using this technique. When generating a 5th order non-linearity, a 3^{rd} order non-linearity is generated as an intermediate step. So it should be possible to generate both non-linear effects within the one circuit and then with appropriate choice of output location the 3^{rd} or 5^{th} order component may be tapped off.

The combined 3^{rd} and 5^{th} order element is shown in figure 5.10. The element operates in the following manner. The input signal is applied to the power splitter where it is split and fed to the LO ports of the first and second mixers. The IF port is fed with an attenuated input signal to prevent saturation of the IF port and the generation of undesirable non-linear components. The RF output of the first mixer is buffer amplified to overcome losses in the mixing process and then split. One of theses signals is used as the LO for the third mixer. The other signal is fed via an attenuator to the IF of the second mixer. The RF output of the second mixer is again buffer amplified to overcome losses and split again. One output provides the 3^{rd} order components for the predistortion process. The second output provides the IF signal for the third mixer. The output of the third mixer provides the 5^{th} order components for the predistortion process.

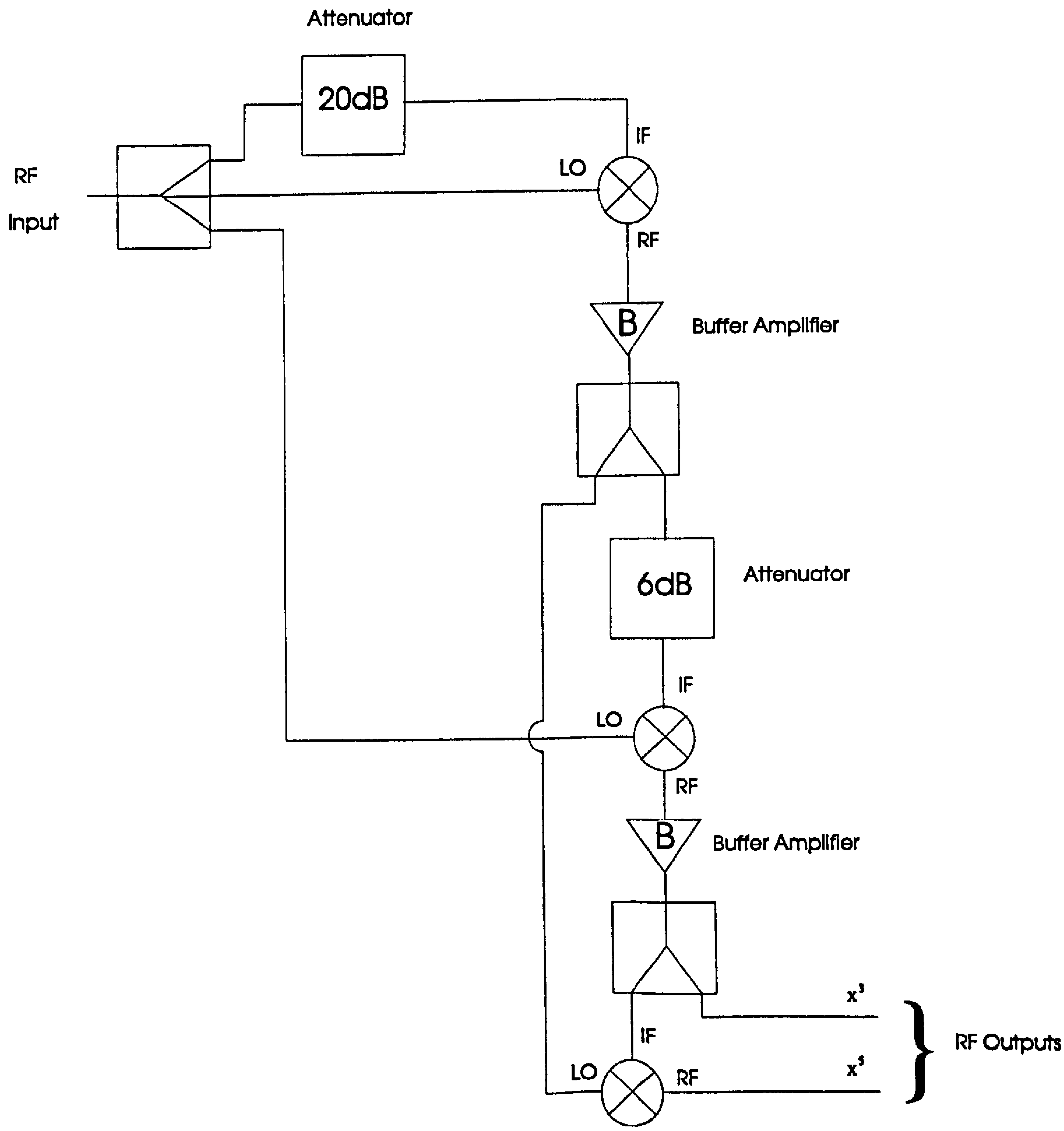

Figure 5.10 Combined 3rd and 5th Order Predistortion Element

5.4.2 3rd and 5th Order Predistortion System

The combined predistortion element was then combined with a complete predistortion system, this system is shown in figure 5.11. The system consists of the combined predistortion element, gain and phase adjustments for 3rd and 5th order products, an attenuator in the main path, time delays in the main and 3rd order predistorted paths, two buffer amplifiers within the predistorter and a buffer amplifier prior to the main amplifier.

The system operates in the following manner. The signal is split at the R.F. input into a main signal and a signal for predistortion. The main signal, which consists of the two main tones, is attenuated and delayed to sum in the correct phase relationship with the signals from the predistorter. The signal which is passed to the predistorter elements is buffer amplified and split equally between the third and fifth order elements. A delay is included in the third order element path to compensate for the extra delay associated with the generation of the 5th order signal. Both the 3rd and 5th order paths have voltage-controlled amplitude and phase control so that the amount of 3rd and 5th order predistortion may be controlled for optimum cancellation. The predistortion signal is then power combined and amplified via a buffer amplifier. This predistorted signal is then combined with the main path signal and buffer amplified before being passed through the main power amplifier and then to the R.F. output.

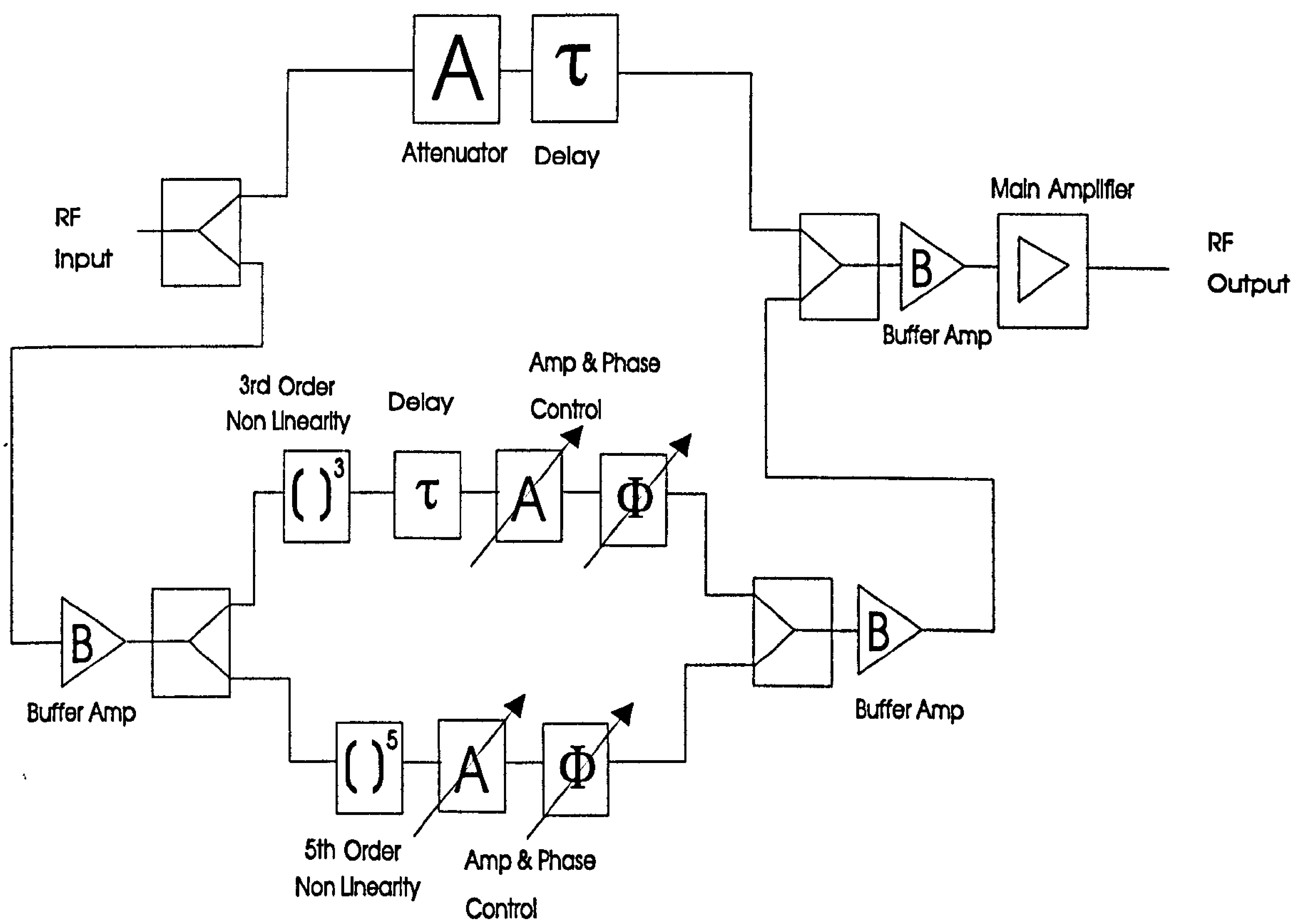

Figure 5.11 Combined 3rd and 5th Order Predistortion System

The circuit in figure 5.11 was used to predistort the Wessex amplifier used in previous investigations. The amplifier output was optimised for the minimum obtainable levels of 3rd order and 5th order intermodulation products. The amplifier was tested at 220MHz with a tone spacing of 500kHz. The results of this predistortion are shown in figure 5.12.

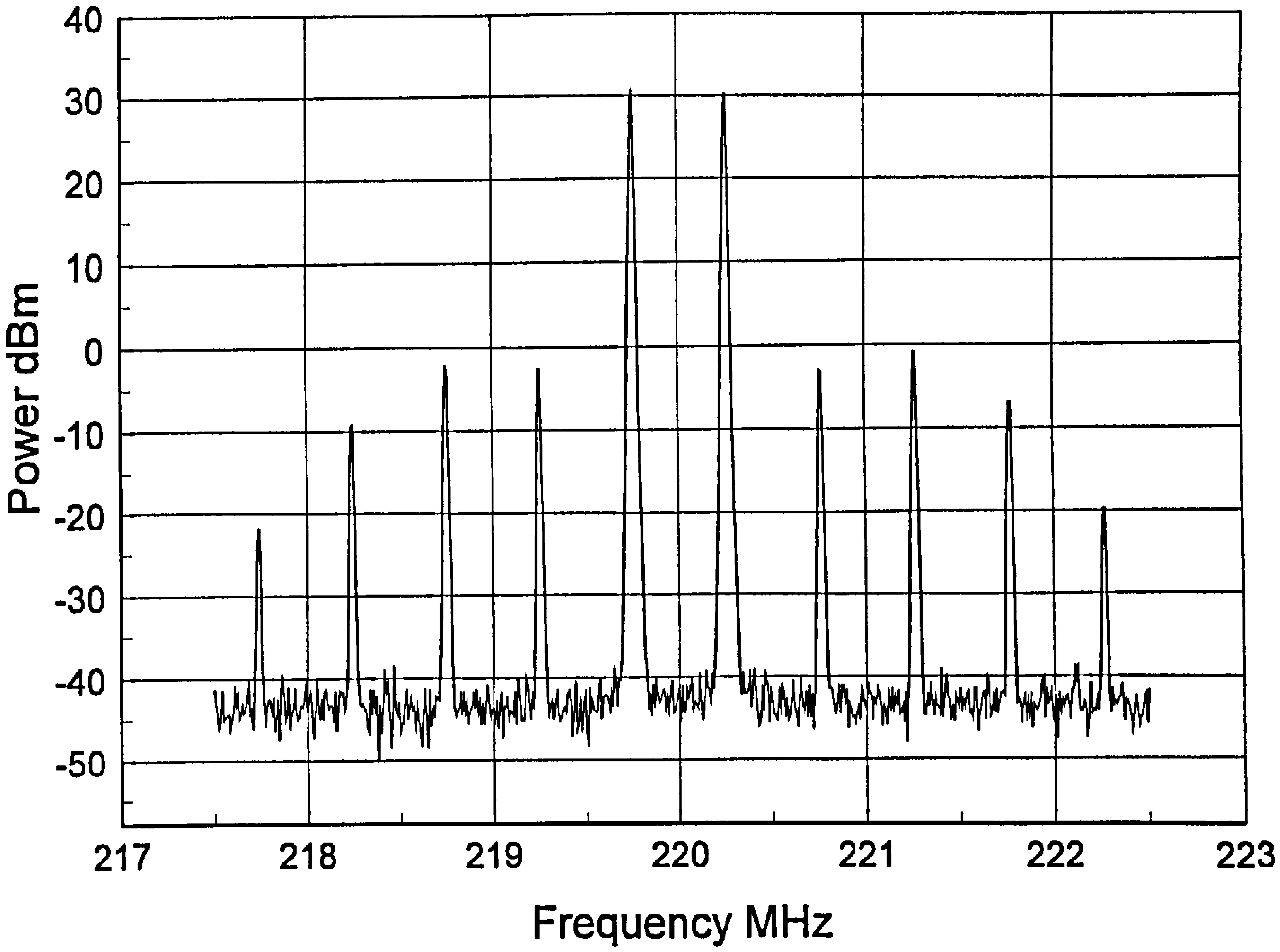

Figure 5.12 Amplifier Output After 3rd and 5th Order Predistortion

Comparison of figures 5.4 and 5.12 show that the 3rd order products are reduced by 10dB, but the 5th order products have increased by 5dB. This leads to the conclusion that 3rd and 5th order predistortion of this type of amplifier does not improve the performance of the Wessex amplifier anymore than 3rd order only predistortion. So the combined predistorter does not provide an improved inverse characteristic for the Wessex amplifier.

The 5th order and combined predistorters do however provide a rich source of intermodulation products and hence it should be possible to predistort a much more non-linear class of power amplifier. The following section describes the use of the 5th order predistorter to linearise a class C power amplifier.

5.5 Predistortion Linearisation of Class C Power Amplifiers

Class C power amplifiers are extremely complex in terms of their modes of operation and the transfer functions that this produces. To produce a practical R.F. or I.F. predistortion system, which could linearise, such an amplifier is unlikely to be a realisable possibility with simple hardware. This means that a class C power amplifier would be better dealt with using a combination of techniques such as predistortion and feedforward. This section deals with the improvement in performance of the class C power amplifier using a predistortion technique.

5.5.1 5th Order Predistortion of Class C Power Amplifiers

The quintic predistorter used in section 5.3 was used to predistort a class C power amplifier which was designed to operate at 220MHz with a peak output power of 36dBm. The amplifier two tone test is shown in figure 5.13. The amplifier performance was measured at 220MHz with a tone spacing of 500kHz. Figure 5.13 shows clearly how the performance of this amplifier is significantly worse than the Wessex amplifier used in previous investigations. The 3rd order IMP's are at –11dBc and it may also be observed that the amplifier exhibits an asymmetric response. The 5th order products are at –18dBc, while the 7th order products are at -28dBc, while the highest 9th order product is at –30dBc. If comparison is made between the Wessex amplifier two tone test in figure 5.4, the Wessex amplifier has a third order performance which is 11dB's better than the class C power amplifier. Also it may be noted that the Wessex amplifier exhibits a symmetrical two-tone test response.

The class C amplifier was predistorted with the quintic predistortion system of figure 5.3, the results of this predistortion are shown in figure 5.14. If figures 5.13 and 5.14 are compared it may be seen that the predistorter has improved the amplifiers performance. The 3rd order IMP's have been reduced by 7dB when compared with the two-tone test. The 5th order products have remained unchanged, the 7th and 9th order products have increased by 3dB and 7dB respectively. This shows how the power has been spread across the band by the predistortion process. The results also show how difficult it is to predistort the much more non-linear classes of power amplifier. In order to see how the predistorter is operating it is necessary to investigate the transfer functions of the amplifier, predistorter and the combined system, the next section covers this investigation.

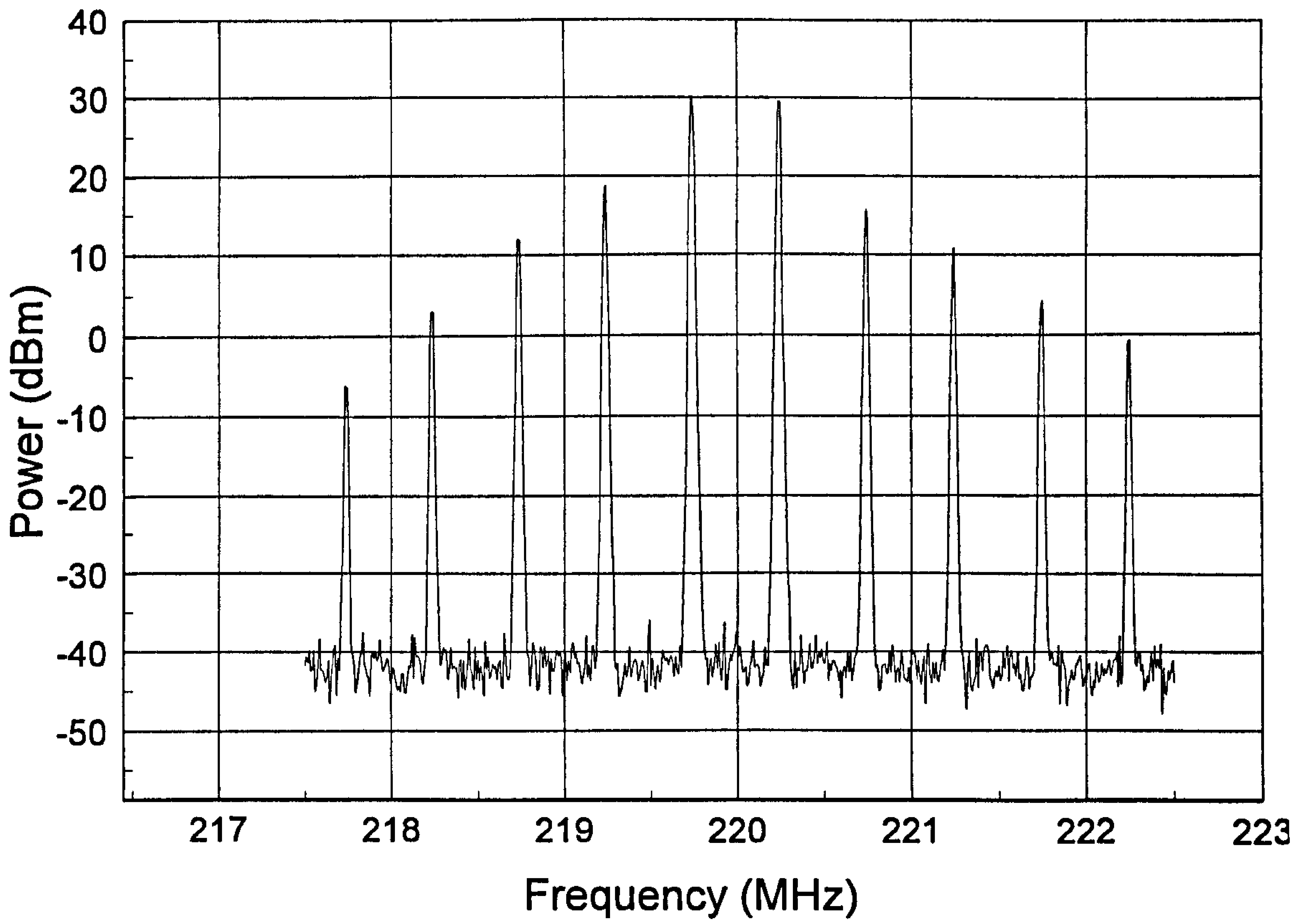

Figure 5.13 Class C Amplifier Two Tone Test at 220MHz with 500kHz Tone Spacing

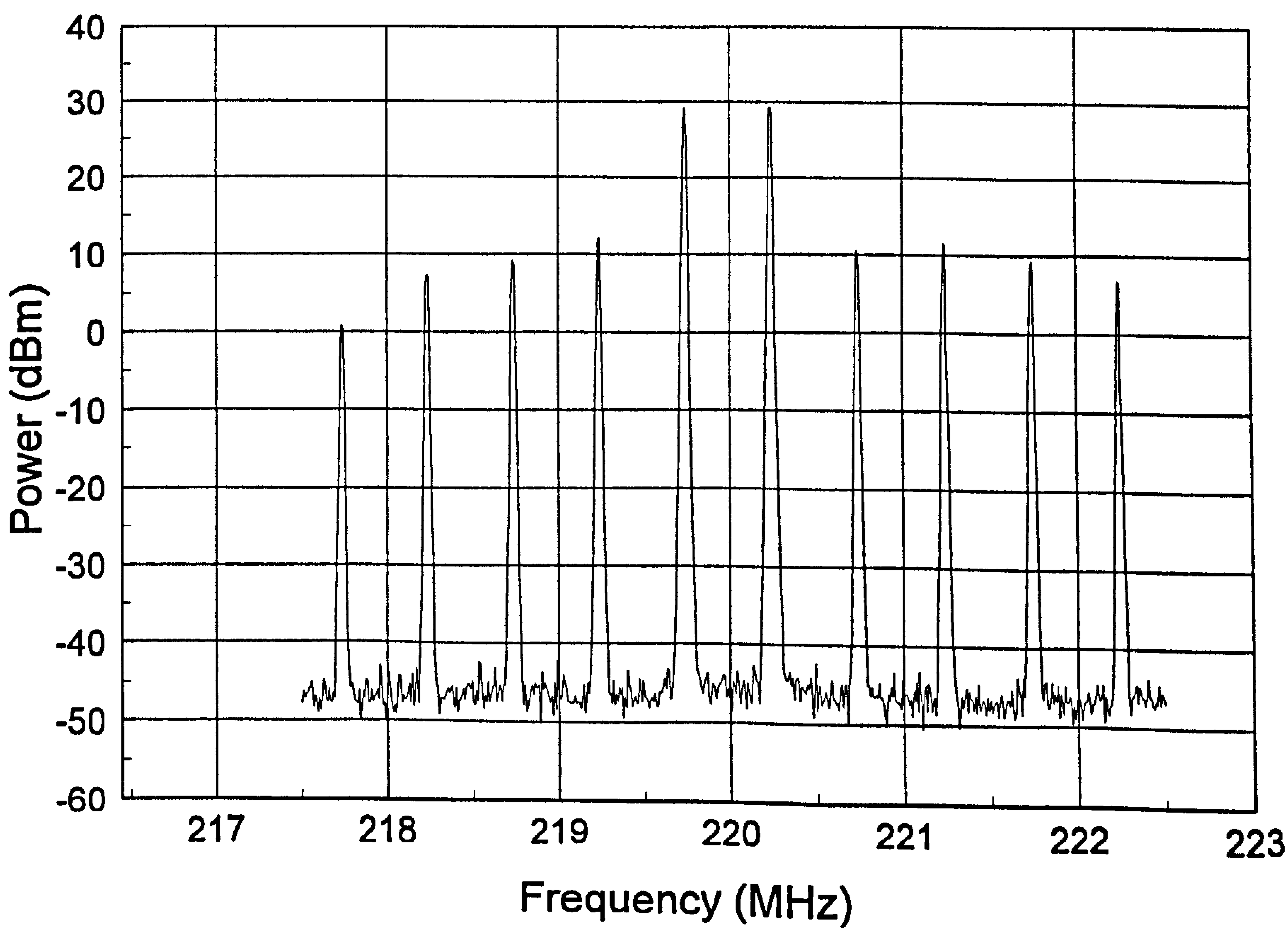

Figure 5.14 Class C Output After 5th Order Predistortion

5.5.2 System Transfer Characteristics

A predistorter should have the exact inverse transfer function of the amplifier being linearised in order for the system to provide the maximum amount of improvement in performance of the amplifier being linearised. This implies that in general a predistortion system can only provide limited improvement in amplifier performance due to the extreme difficulty of generating an exact inverse of the amplifier transfer function. A class C amplifier as stated previously has a complex transfer function. This section investigates the relative matching between the class C power amplifier being predistorted and the predistorter providing the correction.

Class C Amplifier Characteristics

The class C power amplifier that was predistorted in section 5.5.1 has been measured to show its characteristic. The amplifier was measured at 220MHz for gain and phase response, AM to AM and AM to PM response and the I & Q transfer characteristic were measured, these measurements are shown in figures 5.15, 5.16 and 5.17 respectively.

Figure 5.15 shows how the gain varies with input power, reaching a peak of 39dB at an input power of –12dBm and a minimum of 18.5dB at an input power of 4dBm. The phase response also varies greatly with input power level, from an initial value of -180^{o} at –18dBm to a peak value of -90^{o} –3dBm. It may also be observed how the values of gain and phase vary significantly as the input power is varied.

Figure 5.16 shows the AM to AM and the AM to PM transfer characteristics. From figure 5.16 it may be observed how non-linear the AM to AM and the AM to PM characteristics are. Taking the AM to AM characteristic it may be seen that the amplifier has an initial expansive region at levels of input voltage of 0.05V, this is due to the amplifier beginning to conduct as an input voltage is applied. As the input voltage increases the amplifier operates in a more linear mode and then as the voltage rises above 0.15V the amplifier moves into the compressive region of it's characteristic. When the AM to PM characteristic is considered it may be observed that initially the phase change is small with the input voltage less than 0.05V. When the voltage rises above 0.05V significant variation in the AM to PM characteristic occurs. The phase varies by 90^{o} for a variation in input voltage of 0.05 – 0.35V. This transfer characteristic shows the significant amount of distortion that a class C amplifier will add to a signal during amplification.

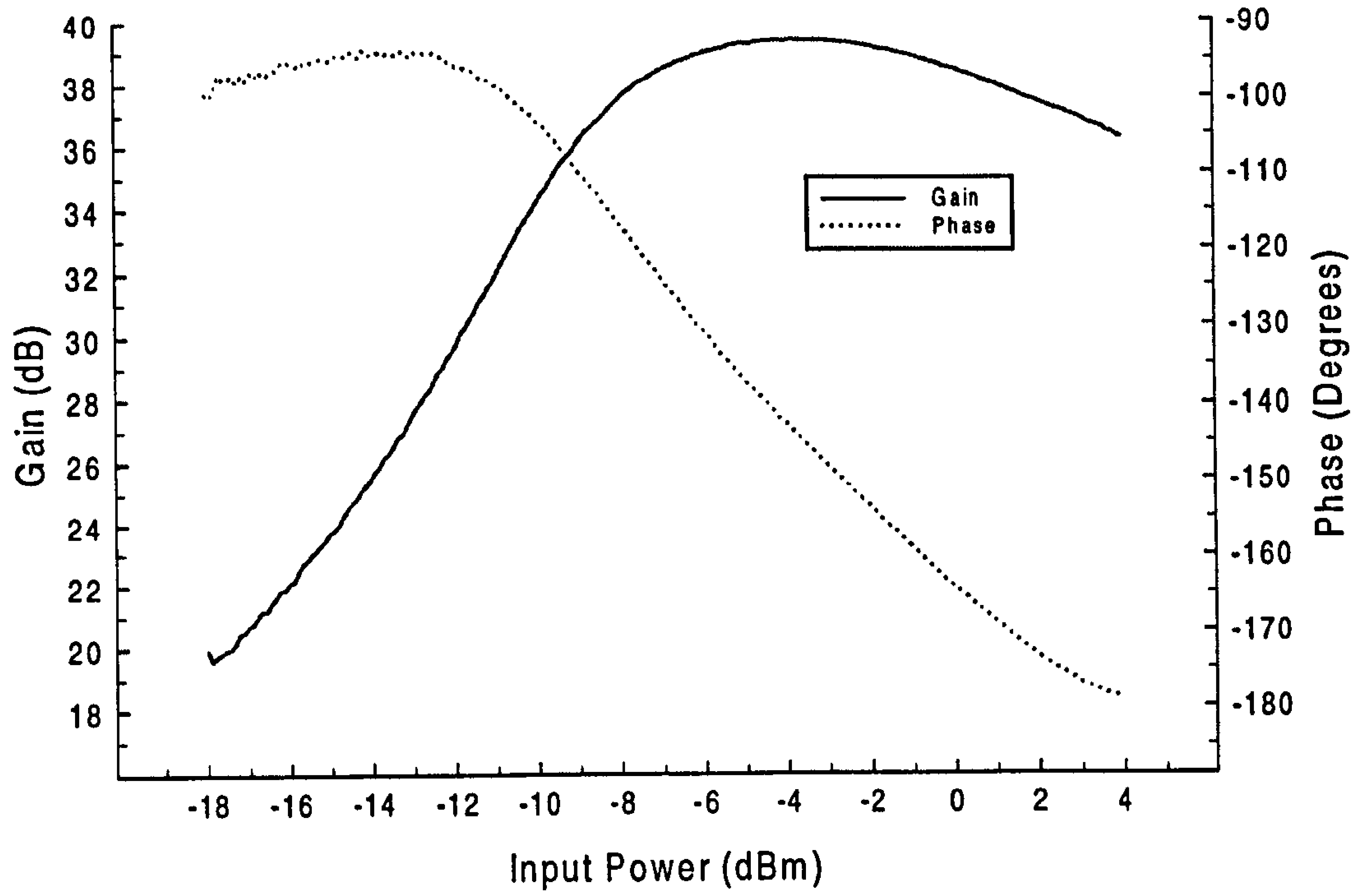

Figure 5.15 Class C Amplifier Gain and Phase Response

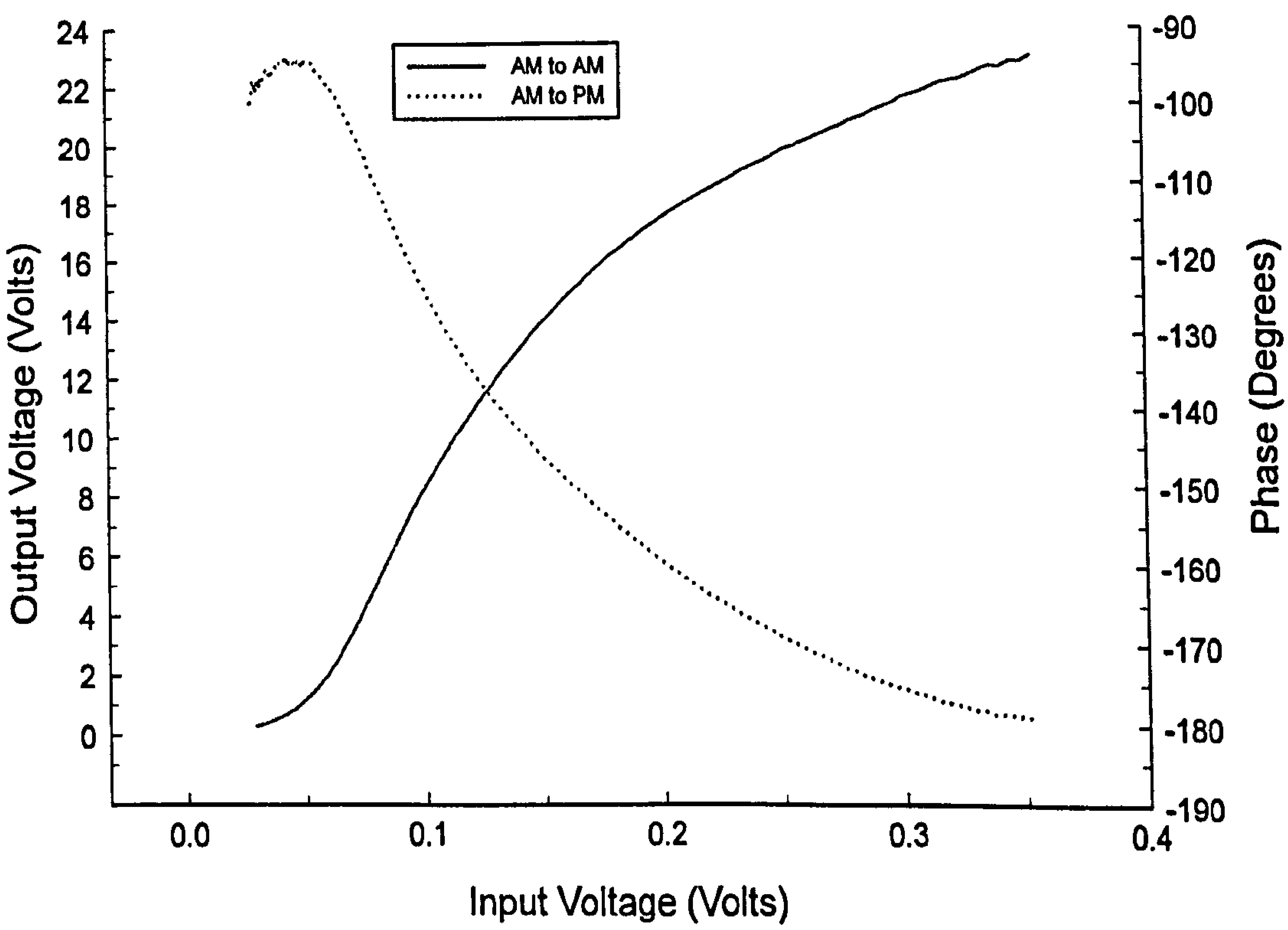

Figure 5.16 Class C Amplifier Voltage Transfer Function

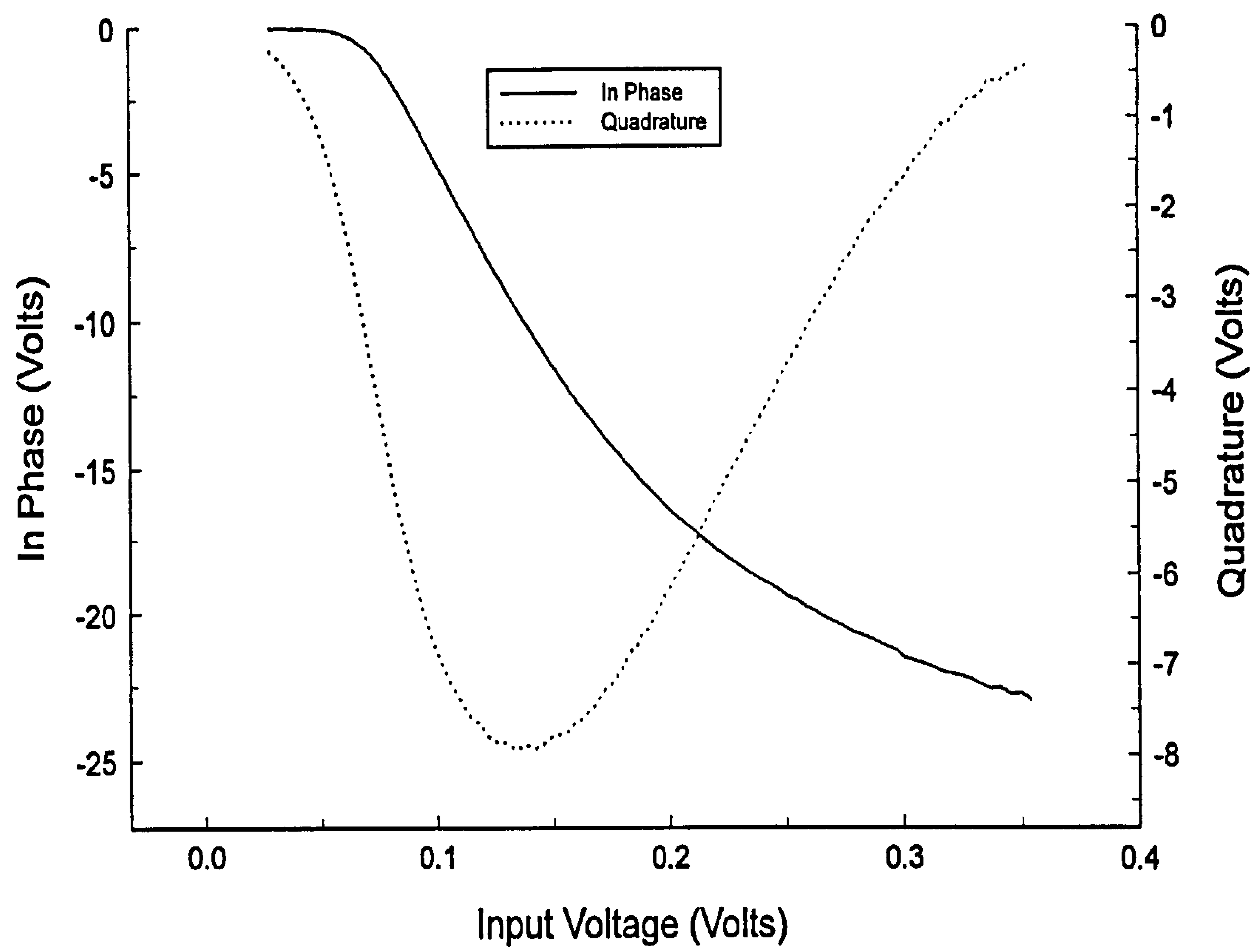

Figure 5.17 Class C Amplifier I & Q Transfer Function

The amplifier's response to In Phase and Quadrature signals is shown in figure 5.17. It may be observed that there is a corresponding effect on the In Phase and Quadrature signals due to the grossly non-linear nature of the amplifier, which has been illustrated by figures 5.15 and 5.16. The amplifier causes the In Phase signal to behave as the inverse of the AM to AM characteristic. The Quadrature signal variation is a combination of distortions due to the AM to AM and the AM to PM characteristics.

5th Order Element Transfer Characteristic

The quintic element which is shown in figure 5.1 had it's gain and phase characteristic, it's voltage transfer characteristic and it's In Phase and Quadrature characteristic measured, these are shown in figures 5.18, 5.19 and 5.20 respectively.

Figure 5.18 shows that the element has an expansive characteristic for input powers in the range of –5dBm to 5dBm. The characteristic becomes compressive from 5dBm to 16dBm. The phase characteristic varies from 100^{o} for an input power of –5dBm to –70 ^{o} for an input power of 0dBm. The phase then varies from -80 ^{o} for an input power of 1dBm to -90 ^{o} for an input power of 16dBm.

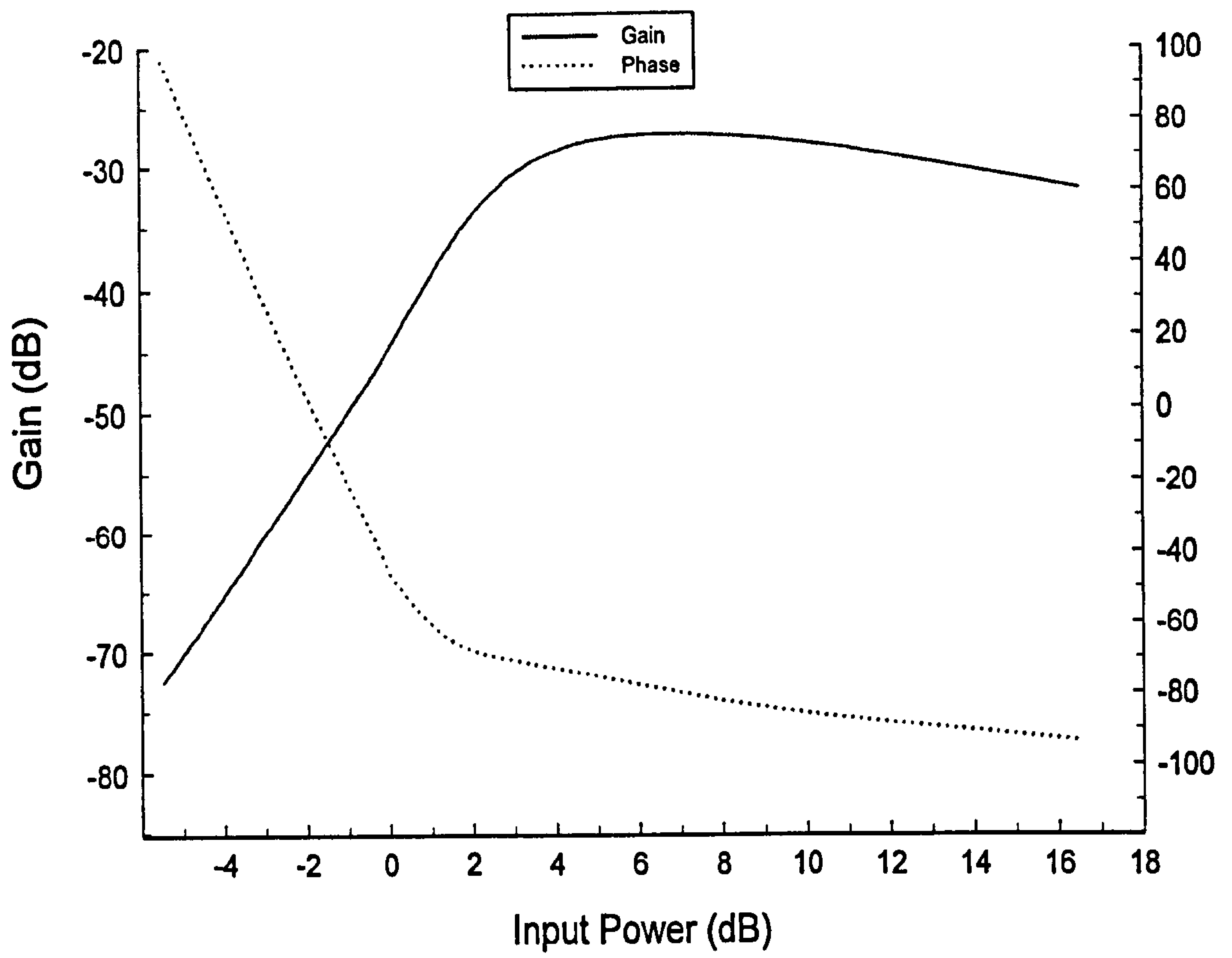

Figure 5.18 5th Order Element Gain and Phase Response

Figure 5.19 shows the voltage transfer characteristic for the 5th order element in terms of the AM to AM and the AM to PM performance. The AM to AM performance of the 5th order element is broadly similar to the AM to AM performance of the class C amplifier as shown in figure 5.16. Figure 5.19 shows how the characteristic is compressive for an input voltage of 0.2 to 0.6V and then becomes extremely compressive for input voltages from 0.7 to 1.4V. The AM to PM characteristic however is not so well matched and exhibits a phase change which is initially very steep and then as input voltage increases becomes progressively more limited.

Figure 5.20 shows the In phase and quadrature performance of the 5th order element. Figure 5.20 shows how the element exhibits a strongly expansive In phase characteristic for values of input voltage less than 0.4V and a strongly compressive characteristic for input voltages greater than 0.4V. The Quadrature characteristic has a strongly compressive characteristic over nearly its whole range. The only exception to this is the initial element turn on at 0.1 – 0.2V.

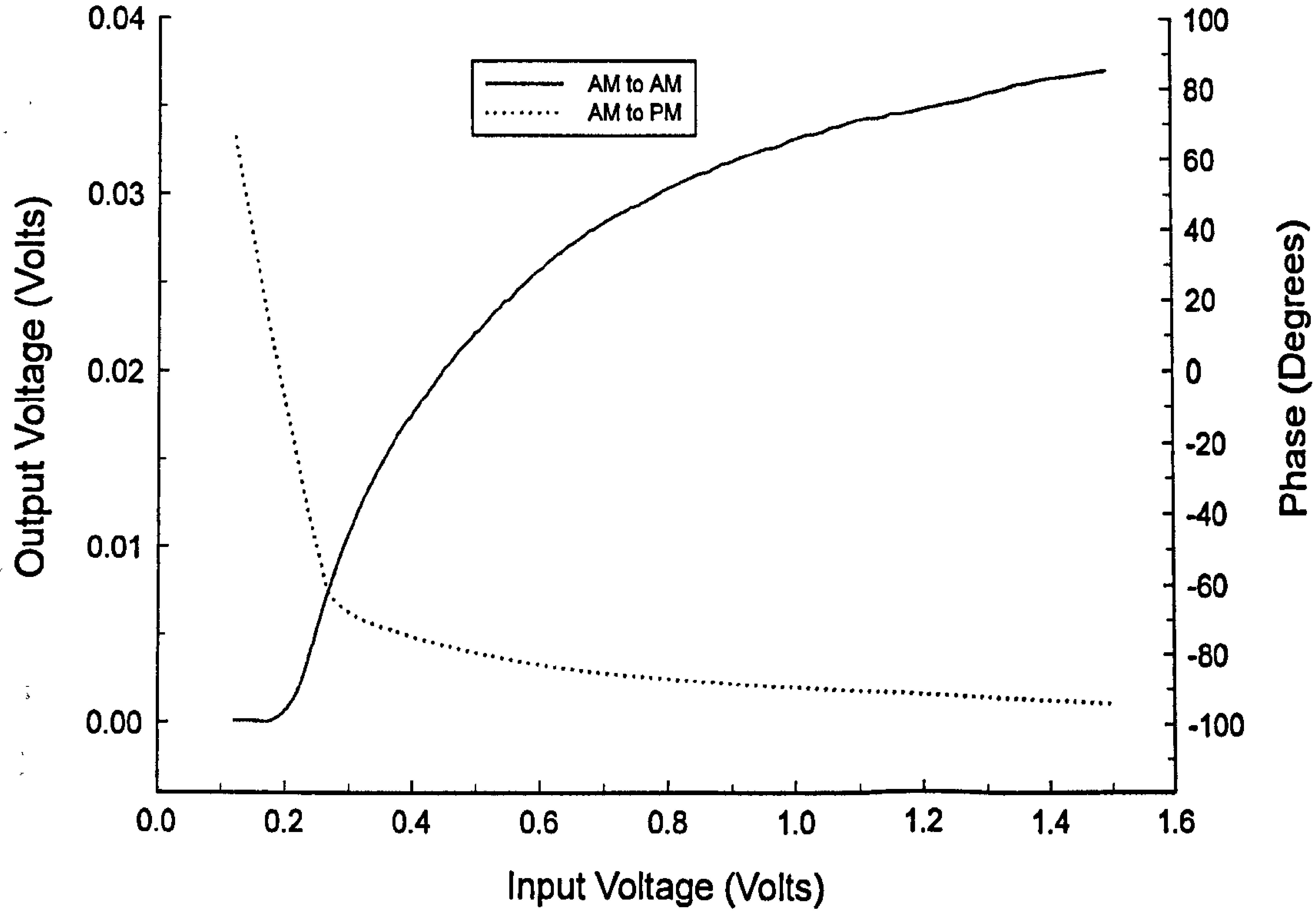

Figure 5.19 5th Order Element Voltage Transfer Function

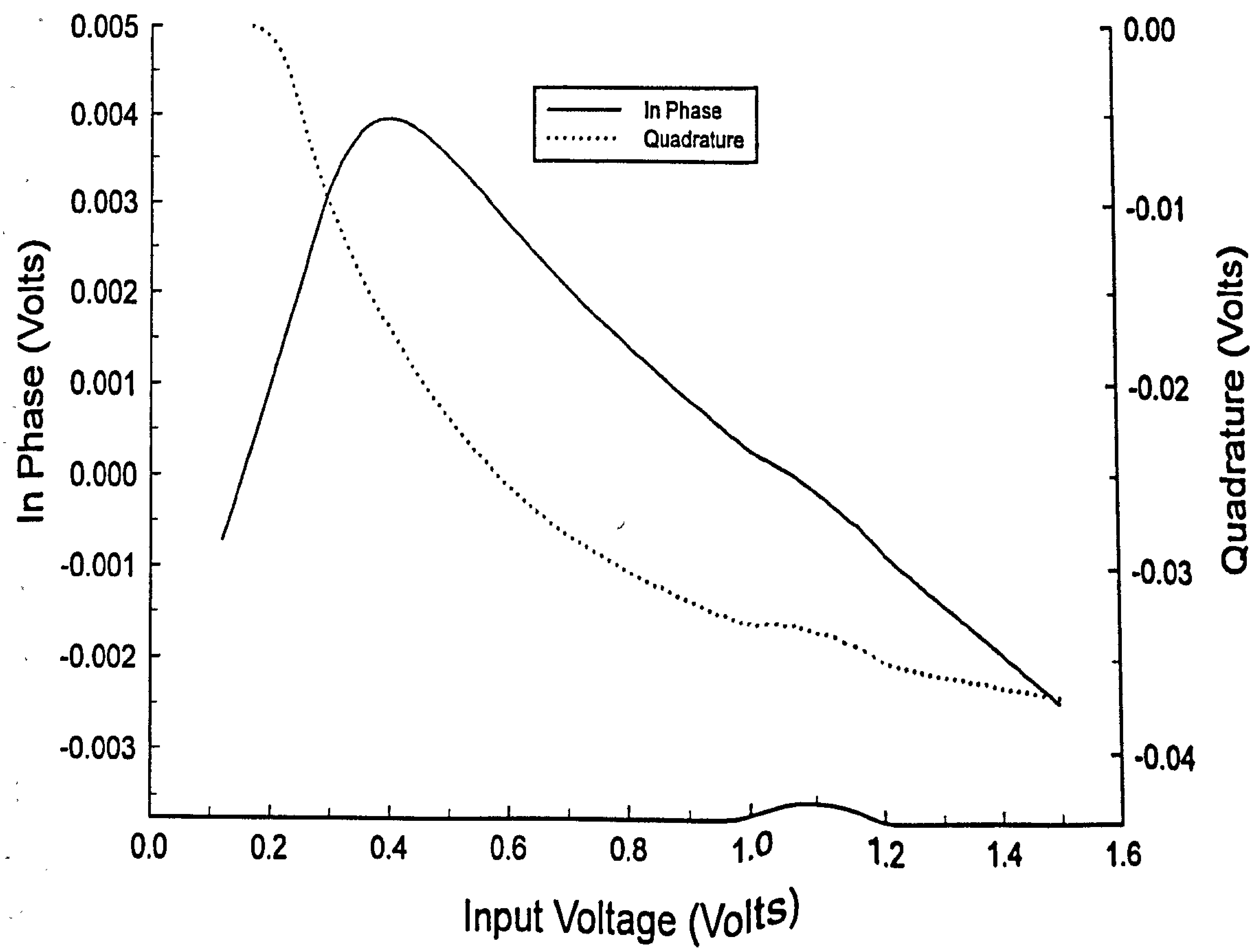

Figure 5.20 5th Order Element I & Q Transfer Characteristic

5^{th} Order Predistorter Transfer Characteristic

The 5^{th} order element of figure 5.1 was then connected into the predistortion system of figure 5.3. The predistortion system gain and phase characteristic, voltage transfer characteristic and the predistorter In phase and Quadrature characteristics were then measured, these measurements are shown in figures 5.21, 5.22 and 5.23 respectively.

Figure 5.21 shows the gain and phase response for the 5^{th} order predistortion system. It can be seen that the system is gain flat for input powers of –5 to 0dBm with a value of gain of –17.7dBm. When the input power rises above 0dBm the gain drops by 4dB for values of input power of 1 to 4dBm. This means that the predistorter has a strongly expansive characteristic in this region. Further falls in gain occur as the input power increases from 4 to 10dBm, this corresponds to a gain fall of 2.5dB. As the input power rises above 10dBm the predistorter exhibits a compressive characteristic with the gain changing from –24.5 to –23dB for input powers of 12 to 16dBm.

The phase of the predistorter is constant for values of input power of –5 to 1dBm. As the input power rises above 1dBm the phase begins to change with the phase being reduced by 42^{O} for input powers of 2 to 8dBm. For input powers of 9 to 16dBm the phase increases by 30^{O}.

Figure 5.22 shows the 5^{th} order predistortion system voltage transfer function in terms of the AM to AM and the AM to PM performance. In general it may be noted that the AM to AM characteristic is initially compressive and then becomes progressively expansive as the input voltage is increased. The AM to AM characteristic is linear for input voltages of 0.1 to 0.3V and then enters a strongly compressive region between 0.3 to 0.4V. The characteristic then becomes expansive for input voltages between 0.4 and 1.5V. The AM to PM characteristic shows a constant phase relationship for input voltages of 0.1 to 0.25V and a phase change of -42^{O} for input voltages of 0.3 to 0.55V. The characteristic then exhibits a positive phase change of 30^{O} for input voltages of 0.6 to 1.5V.

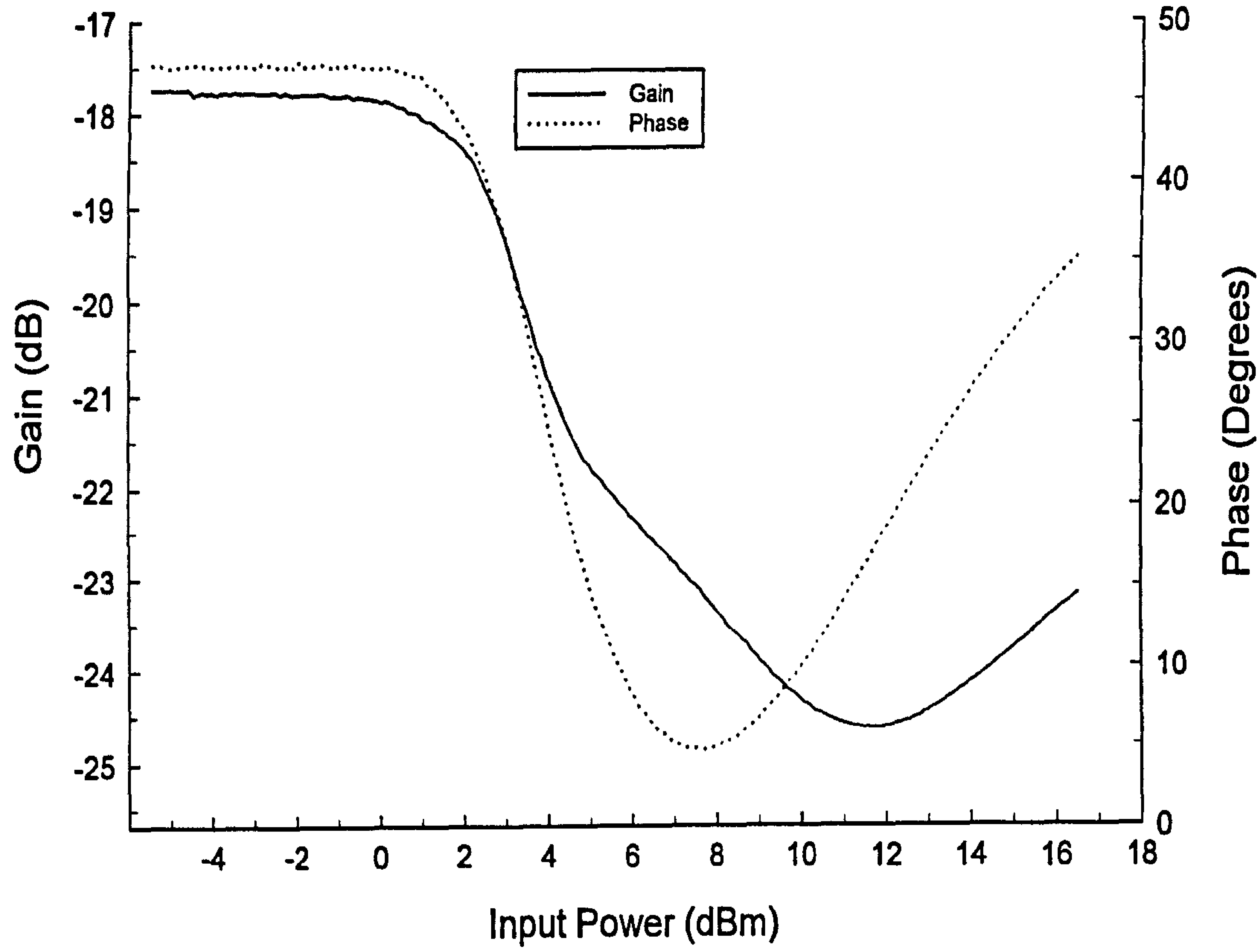

Figure 5.21 5th Order Predistorter Gain and Phase Response

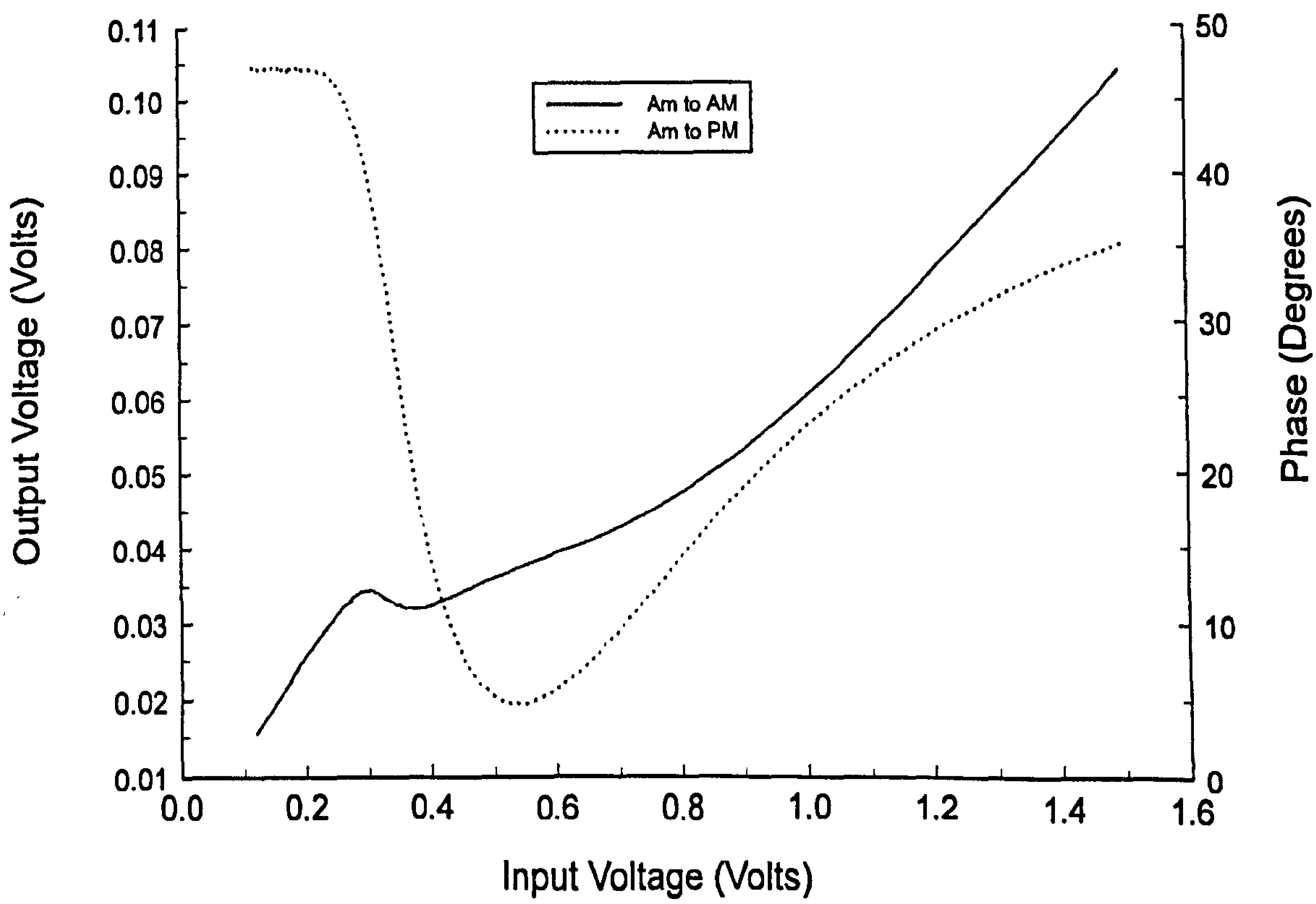

Figure 5.22 5th Order Predistorter Voltage Transfer Function

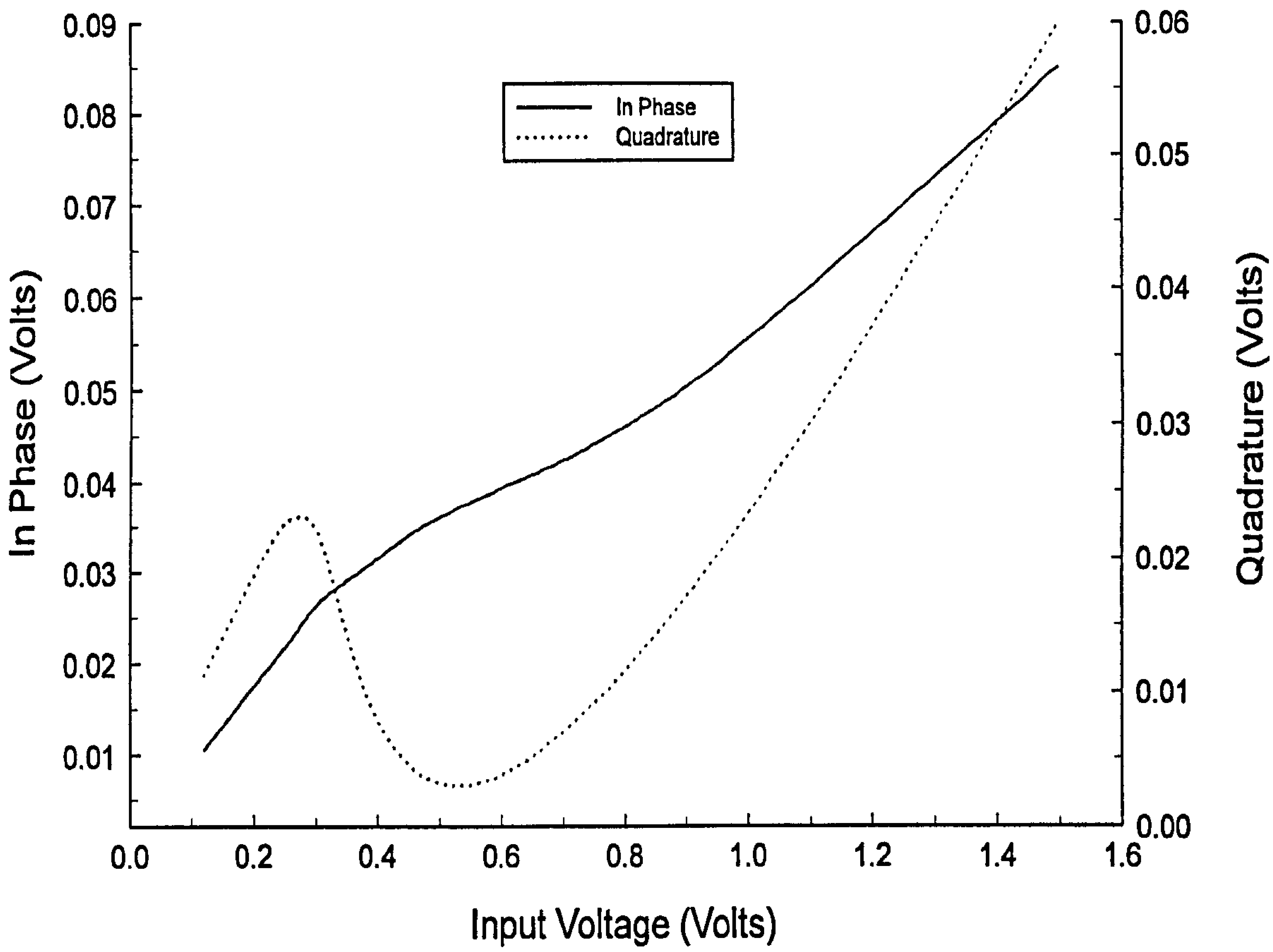

Figure 5.23 5th Order Predistorter I & Q Transfer Function

The 5th order predistorter In phase and Quadrature performance is shown in figure 5.23. The In phase response follows a linear characteristic for input voltages of 0.1 to 0.3V. The characteristic then becomes compressive as the input voltage rises above 0.3V towards 0.5V. As the input voltage rises above 0.6V the characteristic becomes compressive in nature until the voltage reaches 1.1V. For input voltages above 1.1V and up to input voltages of 1.5V the characteristic becomes linear once more. The Quadrature characteristic is linear for values of input voltage from 0.1 to 0.25V. As the input voltage rises above 0.25V the characteristic becomes strongly compressive until the input voltage reaches 0.5V. The characteristic then becomes linear once more as the input voltage rises above 0.6V until the input voltage reaches 1.5V.

Class C Amplifier and Predistortion System Combined Transfer Characteristics

The 5[th] order predistortion system of figure 5.3 was then connected to the class C amplifier that has been characterised previously. Measurements were then taken of the overall system performance. Measurements were taken of the gain and phase performance, the AM to AM and the AM to PM performance and the In phase and Quadrature performance, these results are shown in figures 5.24, 5.25 and 5.26 respectively.

Figure 5.24 shows the overall system gain and phase response. It maybe seen how initially the amplifier system is turned off until the input power rises above 4dBm. The amplification system then exhibits a rapidly changing gain for small changes in input power. This indicates that the amplifier is still far from linear. The system now begins to saturate when the input power rises above 6dBm, strong saturation occurs for input powers above 7.5dBm. The phase response has remained similar in form to the phase response of the original amplifier. The gross phase change has increased however from 90^{O} to 120^{O}.

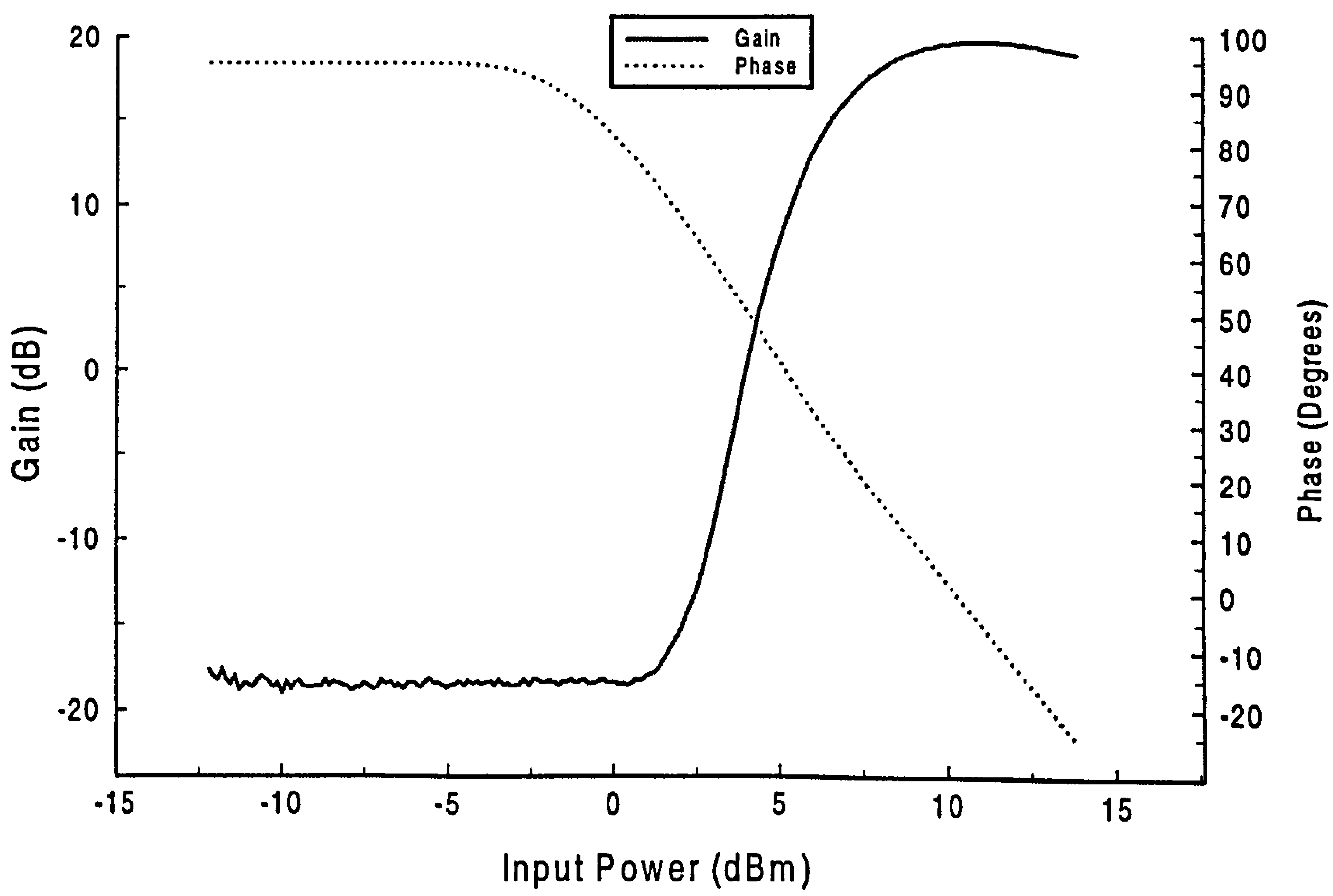

Figure 5.24 Combined System Gain and Phase Response

Figure 5.25 illustrates the system AM to AM and AM to PM transfer characteristic. It maybe observed that the amplifier system is not operational until the input voltage rises above 0.3V. The amplifier system turn on region may be observed as the input voltage changes from 0.35 to 0.38V. The amplifier system has a linear region within the AM to AM characteristic for input voltages of 0.4 to 0.53V. The system then begins to become compressive as the input voltage rises above 0.55V. Compression continues to occur as the input voltage changes from 0.65 to 1.1V. The AM to PM characteristic is of a similar form to the original amplifier characteristic. The phase is initially constant when the system is shut off. As the input voltage rises above 0.2V phase changes begin to occur. As the input voltage changes from 0.22 to 1.1V the corresponding change in phase is 120°.

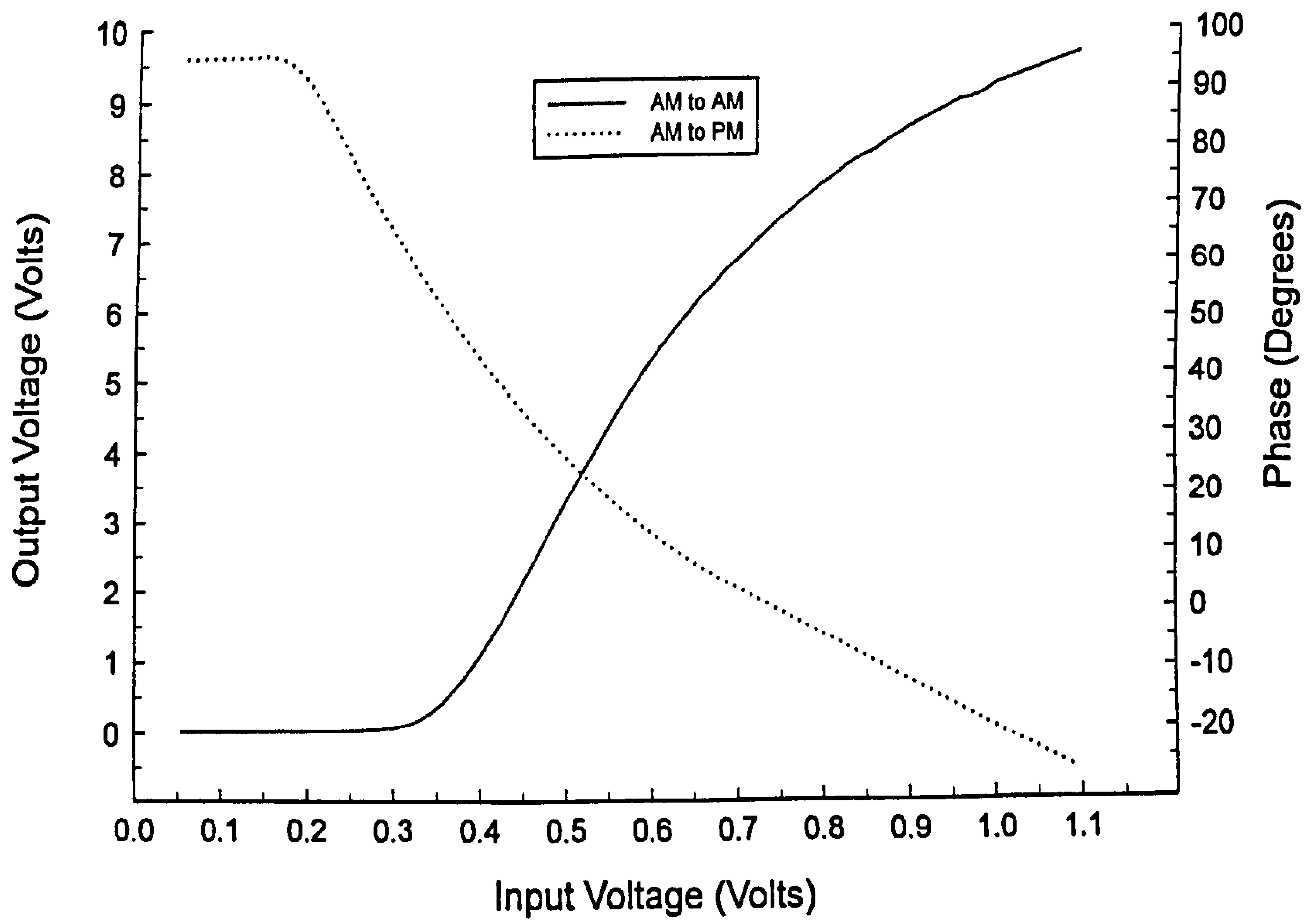

Figure 5.25 Combined System Voltage Transfer Function

Figure 5.26 shows the amplifier system In phase and Quadrature transfer characteristic. The In phase characteristic exhibits little change initially until the system turns on as the input voltage rises above 0.3V. The system turn on occurs as the input voltage rises from 0.32 to 0.42V. Above 0.42V the In Phase characteristic is linear for input voltages from 0.43 to 0.6V. For input voltages of 0.6 to 1.1V the In Phase characteristic becomes increasingly compressive in nature. The Quadrature characteristic also exhibits little change for input voltages less than 0.3V. The Quadrature characteristic has a non-linear response over the whole input range above 0.3V. Initially as the input voltage rises above 0.3V a sharp rise in Quadrature response occurs. This reaches a peak for an input voltage of 0.5V, beyond 0.5V the quadrature voltage falls rapidly until the input voltage reaches a maximum of 1.1V. The non-linear response of the Quadrature component is as a direct result of the large amount of phase change present within the amplifier system.

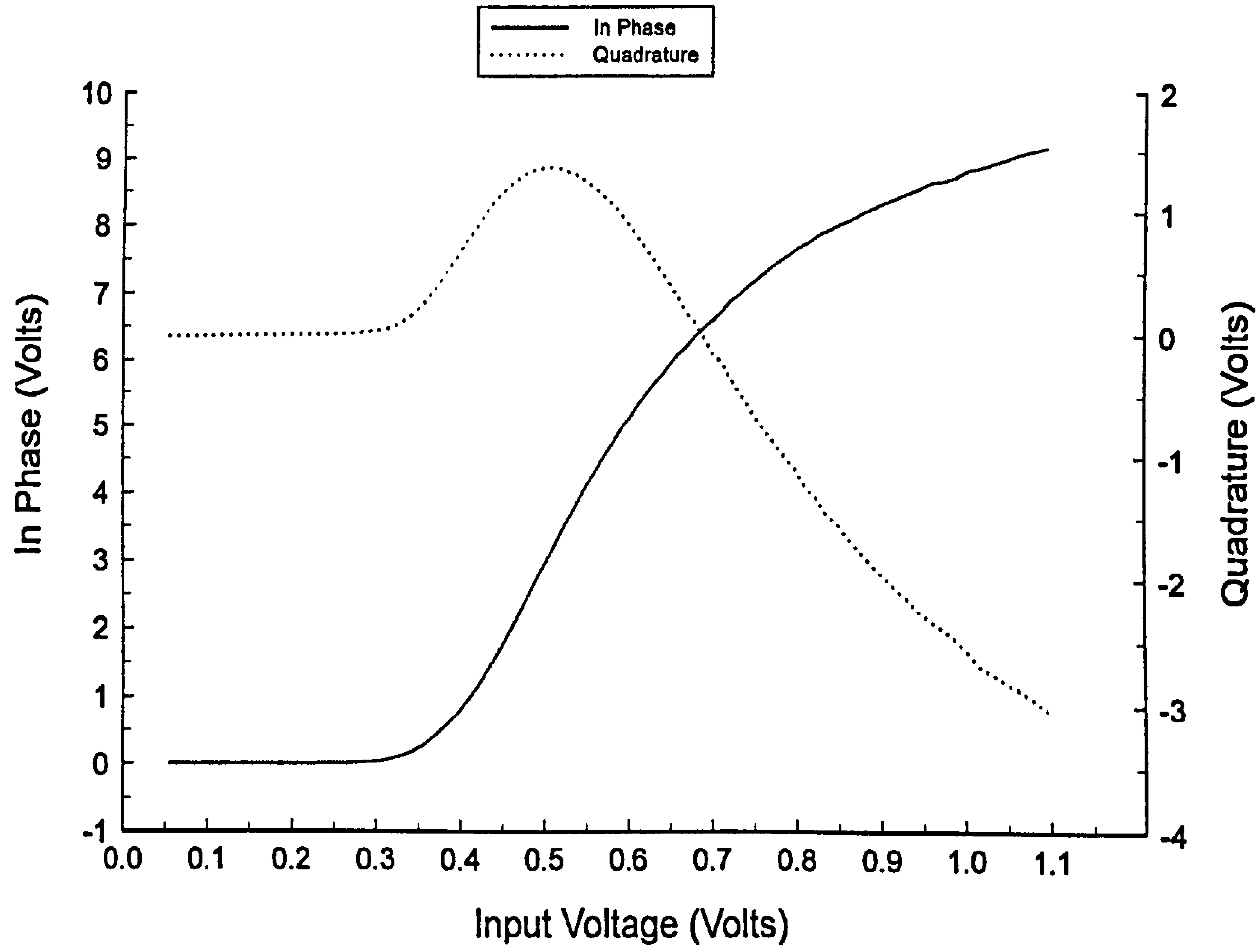

Figure 5.26 Combined System I & Q Transfer Function

5.6 Discussion

The linearisation of any amplifier using predistortion techniques relies on the accurate matching of the amplifier transfer function with an appropriate inverse characteristic generated by the predistortion network. A predistortion system should also be capable of responding to changes in amplifier output power level which will result in a modification in the actual amplifier transfer function, a new predistorter inverse will then be required to compensate for this change in amplifier characteristic.

Initial work carried out with the more linear Wessex amplifier has shown that it is possible to reduce the 5th order IMP's independently of the other products. Therefore it is theoretically possible to control independently the 3rd and 5th order IMP's. However when both 3rd and 5th order compensation is applied then the amplifier performance is degraded when compared with 3rd order compensation alone. This indicates that the 3rd order predistorter provides a much better inverse to the amplifier characteristic than the combined 3rd and 5th order system.

When considering the linearisation of class C amplifiers the problems are considerably more difficult to solve due to the much more complex nature of the amplifier characteristic. This additional complexity is due to the class C technique requiring the transistor to operate over the whole of it's characteristic, from cut-off right through to saturation during one waveform cycle. The lineariser also has to contend with an amplifier, which has a large amount of AM to PM distortion in addition to the AM to AM distortion. The Wessex amplifier, which has been linearised within chapter 4, has a characteristic which contains mainly AM to AM distortion and very little AM to PM distortion. This accounts partially for the better performance of the lineariser when used with the Wessex amplifier. An additional effect, which is likely to have a significant effect on the overall system performance, is the frequency dependency that an amplifier will exhibit due to the characteristics of the device used within the amplifier. The effect of frequency dependency[1] may be observed by measuring the amplifier transfer characteristic at various frequencies [4]. This effect is also in part responsible for poor intermodulation performance of class C amplifiers. The reason for this is that, the two-tone test measures amplifier performance over a band of frequencies. Several models have been suggested over the years [5, 6] to try and quantify frequency dependency. These models are concerned with the behaviour of TWT amplifiers which are generally more linear than class C amplifiers and so none of these models has fully explained or described how a highly non-linear amplifier does behave. It has been proposed in [4] that piecewise

[1] Often referred to as amplification with memory.

linear predistortion may be used to linearise class C amplifiers. Extensive simulations were conducted to verify the practicality of this approach. However the simulations had to operate under a number of assumptions for computational reasons, the main one being that the class C amplifier being linearised was frequency independent. When a practical predistortion system is constructed then this assumption is no longer valid. So it would appear that a predistortion system that is required to linearise a class C amplifier will need to be capable of varying it's characteristic with frequency as well as with variations in amplitude if really useful improvements in linearity are to be achieved.

5.7 Summary

This chapter has introduced the development of a quintic polynomial system for the broadband linearisation of R.F. power amplifiers. It has been shown that it is possible to reduce the 5^{th} order products of an R.F. power amplifier using quintic predistortion techniques. The use of predistortion is dependent on the ability of the network to accurately map the inverse characteristic of the amplifier being linearised. This chapter has shown that it is also possible to achieve limited improvement in the performance of class C amplifiers with the use of quintic predistortion techniques. The performance of a predistortion system is limited by the ability of the network to accurately match the extremely complex characteristics of highly non-linear amplifiers. Performance is also limited by the considerable difficulty in producing a network that is capable of matching the changes in characteristic caused by the frequency dependency of these extremely non-linear amplifiers.

References

[1] M.B. Steer and P.J. Khan, An algebraic formula for the output of a system with large-signal, multifrequency excitation, Proceedings of the IEEE Vol. 71, No. 1, pp 177 – 179, January 1983.

[2] G.W. Rhyne, M.B. Steer and B.D. Bates, Frequency-domain nonlinear circuit analysis using generalized power series, IEEE Transactions on Microwave Theory and Techniques, Vol. MTT 36, No. 2, pp 379 – 387, February 1988.

[3] Mini Circuits RF/IF designers handbook, 1997.

[4] K. P. Parsons, Highly Efficient Broadband Linearisation of R. F. Amplifiers for Satellite and Cellular Applications, University of Bristol, March 1996.

[5] A.A.M. Saleh, Frequency-independent and frequency-dependent nonlinear models of TWT amplifiers, IEEE Transactions on Communications, Vol. COM 34, No. 1, pp 67 – 71, January 1986.

[6] M.T. Abuelma'atti, Frequency-dependent nonlinear quadrature model of TWT amplifiers, IEEE Transactions on Communications, Vol. COM 29, No. 11, pp 982 – 986, August 1984.

Chapter 6
Conclusions and Suggestions for Further Work

6 Conclusions and Suggestions for Further Work

6.1 Amplifiers for Software Radios

Amplifier linearity has always been a problem in communications systems. With the advent of standards such as UMTS with ever increasing bandwidth and linearity requirements this issue is becoming increasingly important. Operators are naturally reluctant to replace expensive hardware when a new standard is implemented due to the expense involved. This has spurred the development of the software radio concept. A key enabling technology of the software radio is the linear power amplifier. Linear amplifiers will be necessary in both the basestation and the mobile sections of any software radio system. It has been shown that no single method of linearisation can provide the solution to the problems, which are now faced by network operators and equipment manufactures in terms of the software radio concept. Methods such as CALLUM can potentially be used within the mobile sections of the system. However due to the limited bandwidth of such amplifier topologies these methods are totally unsuitable for use within the basestation system as MCPA's. The only viable solutions to the basestation problem are feedforward and predistortion due to the broadband capability of such systems. This work has shown that there is almost no limit to the bandwidth that may be achieved by an analogue predistortion system.

The solution to the problem of high efficiency and high linearity broadband operation is to be found in hybrid techniques. Predistortion is one of the most efficient forms of broadband linearisation available. However predistortion cannot provide the levels of linearity required by modern communications systems. Feedforward can provide the required levels of linearity but the efficiency of feedforward systems is generally very poor. Feedforward systems are fundamentally limited in their performance due to the need for two power amplifiers, the main amplifier and the error amplifier. To boost the efficiency of feedforward systems as much as possible it is desirable to use as non-linear an amplifier as possible as the main amplifier. This places a high power requirement on the error amplifier. If the main amplifier can be partially linearised by techniques which have minimal impact on the amplifiers efficiency then the required error amplifiers power handling requirement can be reduced. Predistortion is ideally suited to the task of limited linearisation with minimal impact on the efficiency of the system. For this reason this thesis has investigated the capabilities of analogue predistortion systems.

6.2 Polynomial Predistortion

Polynomial predistortion approximately models an amplifier as a set of polynomial functions. It has been shown that the potential gains in linearity of a 3rd order amplifier are considerable when correct gain and phase control are applied. The analysis within this thesis has shown how the optimum values of 3rd order intermodulation products may be calculated. This analysis was then used to investigate the effect of gain and phase errors on cancellation performance. These investigations show that the location of the gain and phase adjustment has an effect on the performance of the system. There is a 10dB improvement in performance if the gain and phase control is applied after the cubic no-linearity. The results show that very tight gain and phase matching is required if optimum performance is to be achieved. The results show quite clearly that the predistortion system is very intolerant of gain error indeed. The simulations have shown that if the amplitude error exceeds 0.5dB then the phase has no effect on the performance of the system until the phase error exceeds 2^{o}. This level of error is very likely within a practical system. The simulations also show that when an amplifier is predistorted using a cubic predistorter while 3rd order products are reduced the 5th order products increase. This means that complete cancellation of the 3rd order products is actually undesirable because the level of the 5th order products then dictates the amplifiers performance. So for a practical system cancellation of the 3rd order products below the level of the 5th order products is unnecessary. This places a limit on the required performance of the predistortion system gain and phase matching requirements of 0.1dB amplitude error and 0.5^{o} phase error.

6.3 Practical Cubic Predistortion

It has been shown that extremely broadband improvements in amplifier performance are possible using the cubic predistortion system within this thesis. With an instantaneous bandwidth of up to 180MHz this predistorter has a bandwidth beyond any predistortion system previously reported. The system is capable of providing 25dB of 3rd order cancellation, this is extremely impressive when it is considered that the predistortion system can operate over a decade of frequencies while still providing significant improvements in amplifier performance. This makes this form of predistortion well suited to software radio applications. All components used in the design of the cubic predistortion system are available over a range of frequencies. So this technique could be easily adapted to operate at the frequencies currently used by second generation systems and the currently planned frequencies for third generation.

It has been shown that careful design of the system is necessary if optimum performance is to be achieved. The balancing of delay between the predistorted path and the main path is critical if optimum cancellation is to be achieved. The choice of component used within the predistortion element has an effect on the actual level of performance. The use of mixers within the predistortion system has been shown to yield significant benefits in terms of the flexibility of the system. The system may be used over any range of frequencies where a suitable mixer may be obtained. The choice of mixers as the multiplying block within the system does however introduce undesirable additional intermodulation products that degrade the system performance to a limited degree.

The results in chapter 4 show that the multiplier system can provide 5dB more 3^{rd} order cancellation than the mixer-based system. Also the 5^{th} order products are reduced by 10dB, the 7^{th} order products are eliminated and no additional products are generated. However multiplier technology currently is not capable of operating at the frequencies required by mobile cellular networks, with advances in device technology this may change making multiplier predistorters an attractive form of amplifier linearisation. Multiplier linearisers also have more limited bandwidth capability than the mixer-based systems.

6.4 Higher Order Polynomial Predistortion

It has been shown that it is possible to reduce the 5^{th} order products of an amplifier using a quintic predistortion system. The amplifier chosen for this investigation was predominantly third order non-linear with only a limited amount of 5^{th} order non-linearity. When a combined 3^{rd} and 5^{th} order predistorter was used with this amplifier the resultant predistorter transfer function was actually a poorer inverse transfer function than the 3^{rd} order predistortion system. This resulted in degradation in the performance of the overall system.

Class C amplifiers are extremely non-linear in terms of their AM to AM and AM to PM performance. It is possible to a limited degree to linearise these amplifiers. It has been shown that the class C amplifiers performance may be improved by 7dB using a quintic predistortion technique. Predistortion has a minimal effect on the efficiency of the amplifier system and so even this limited amount of improvement is useful if predistortion is combined with another linearisation technique such as feedforward. If a combined amplifier were to be used in this case the overall efficiency would be almost 38% which is an 11% improvement when compared with the use of the feedforward in isolation, this analysis is shown in appendix A. The mixer-based lineariser has a characteristic, which does not inversely match closely enough the characteristic of the class C amplifier. Also the predistortion system does not change it's characteristic sufficiently over the operating bandwidth to compensate for the frequency dependency of the class C amplifier.

6.5 Concluding Remarks

This thesis has shown that within the near future linear power amplifiers are going to be essential for the operation of future systems. Software radio is likely to play an important part in the future of communications systems. Current second-generation systems such as GSM are likely to be in use for at least the next 10 years. Even though current standardisation activity was aimed at a single global standard, the compromises that were agreed for political and commercial reasons have resulted in a group of standards. GSM EDGE is now designed to evolve over time, this means that a reconfigurable basestation is likely to be very desirable.

Predistortion is an extremely attractive method for the efficient broadband linearisation of power amplifiers. This work has shown that R.F. polynomial predistortion can generate useful improvements in amplifier performance. However for amplifiers with complex transfer functions such as class C power amplifiers only very limited improvements can be achieved. So predistortion needs to be used with techniques such as feedforward to obtain the required levels of linearity.

6.6 Suggestions for Further Research

This thesis has shown that cubic predistortion can provide very useful improvements in amplifier performance. Little research has been carried out on the control of R.F. predistortion systems. This is an area that requires investigation in terms of the complexity of such control schemes. The cubic technique developed within this thesis is extremely simple and to make it attractive as a practical technique an equally simple control scheme would be required.

This thesis has investigated scalar predistortion systems which operate on the I & Q signals simultaneously. It has been shown in this thesis that it would be desirable to independently alter the I & Q signals. It would be instructive to extend the techniques investigated within this thesis to independently control the amount of I & Q predistortion applied to the amplifier. This is especially applicable to the linearisation of class C power amplifiers where a significant amount of additional Quadrature distortion is present.

The cubic predistortion system that has been developed could be applied to other parts of the radio system that require linearisation. This system would be particularly attractive to the areas of the radio, which are predominately 3rd order non-linear such as the mixers and modulators. Very limited research has been carried out in this area [1] but if truly broadband reconfigurable radio systems are to be developed then this will be an area which creates severe restrictions on what can be achieved in terms of the overall linearity of the complete radio system.

References

[1] Correction of Mixer Nonlinearity in Quadrature Modulators, by M. Faulkner and M. Johansson, in Electronics Letters, January 1992 Vol. 28 No. 3 pp. 293–295.

Appendix A

A.1 Analysis of Efficiency Improvements for Predistortion and Feedforward Hybrid Amplification System

This appendix carries out an analysis of the potential efficiency gains, which are possible by combining feedforward and predistortion into a hybrid amplifier structure [1].

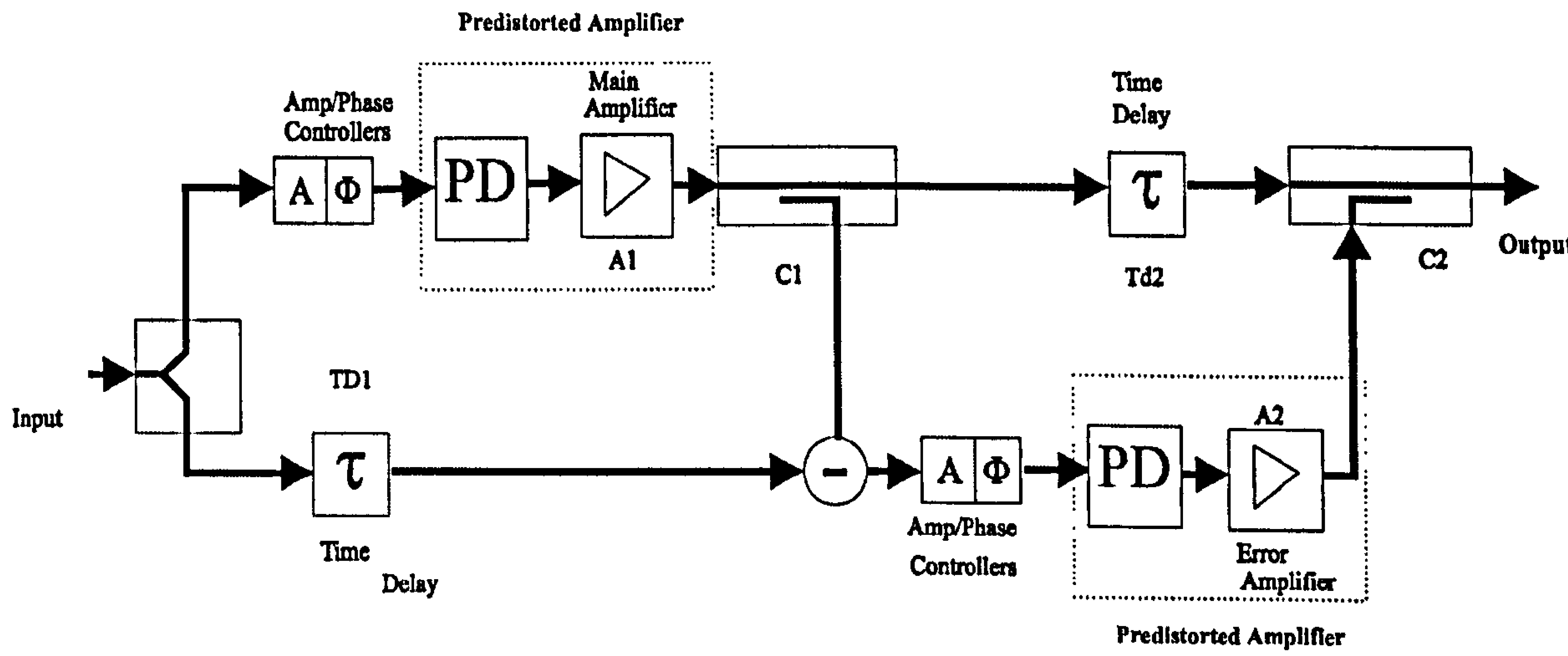

Figure A.1 Predistortion and Feedforward Amplifier

Using the architecture shown in figure A1 it has been shown by Parsons [2] that the efficiency of a practical feedforward amplifier may be expressed as:

$$\eta_{ff} = \frac{\eta_{A1}\eta_{A2}C_{DC}(1-C_{DC})}{\eta_{A1}F_{IM}(1-C_{DC})+\eta_{A2}LC_{DC}(1+F_{IM})} \qquad \text{(A.1)}$$

Where: η_{A1} is the main amplifier efficiency

η_{A2} is the error amplifier efficiency

C_{DC} is the output coupler coupling factor

F_{IM} is the fractional power of a single IMP

L is the fractional loss of signals through the delay

It has also been shown in [2] that there is an optimum value of coupling factor $C_{DC,OPT}$ which is given by:

$$C_{DC,OPT} = \frac{\eta_{A1}F_{IM} \pm \sqrt{\eta_{A1}\eta_{A2}F_{IM}L(1+F_{IM})}}{\eta_{A1}F_{IM} - \eta_{A2}L(1+F_{IM})} \qquad \text{(A.2)}$$

It has been shown in chapter 5 that it is possible to improve the linearity performance of a class C power amplifier by the use of quintic predistortion techniques. It has been shown that the performance of a class C amplifier may be improved by 7dB. Using this figure, calculations may be carried out to find the expected efficiency of the system.

Assuming that:

$\eta_{A1} = 60\%$

$\eta_{A2} = 60\%$

L = 1dB

If the amplifier has a raw uncorrected performance of – 11dBc then:

$$F_{IM} = \text{-18dBc} \qquad \text{(A.3)}$$

Therefore:

$$C_{DC,OPT} = 8.98dB \qquad \text{(A.4)}$$

Using this value of $C_{DC,OPT}$ gives:

$$\eta_{ff} = 37.75\% \qquad \text{(A.5)}$$

So with 7dB of predistortion cancellation it is possible to improve the efficiency of the feedforward amplifier by 11%.

References

[1] P. B. Kenington, K. J. Parsons and D. W. Bennett, Broadband linearisation of high-efficiency power amplifiers, Proceedings of the Third International Mobile Satellite Conference, pages 59 – 64, June 1993.

[2] K. J. Parsons and P. B. Kenington, The Efficiency of a Feedforward Amplifier with Delay Loss, IEEE Transactions on Vehicular Technology, Vol 43, No. 2, pages 407 – 412, May 1994.